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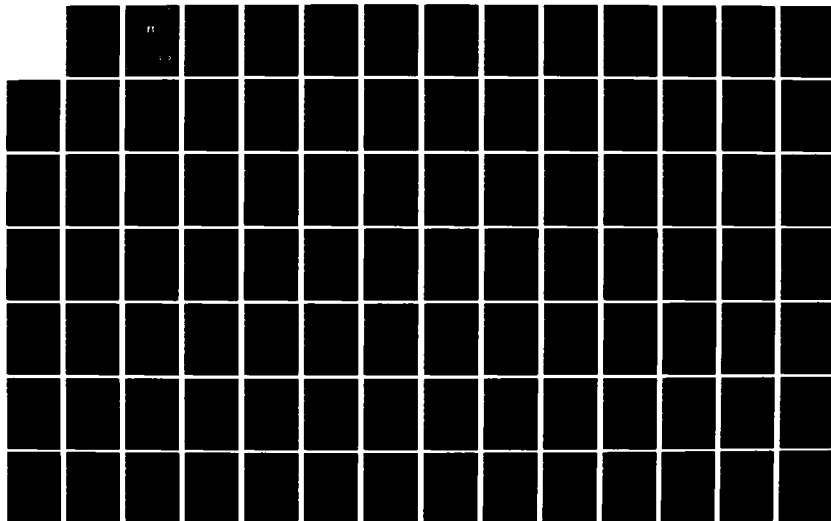
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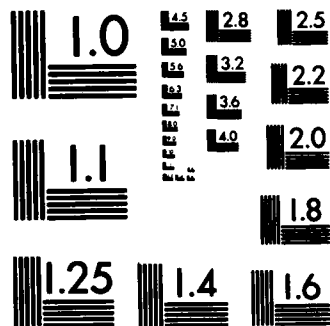
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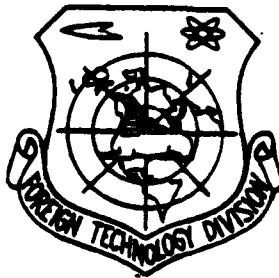
FOREIGN TECHNOLOGY DIVISION



FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC BAND

by

V.M. Volkov



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FOREIGN TECHNOLOGY DIVISION
WP-AFB, OHIO.



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U. S. BOARD ON GEOGRAPHIC NAMES TRANSLITERATION SYSTEM

Block	Italic	Transliteration	Block	Italic	Transliteration
А а	<i>А а</i>	A, a	Р р	<i>Р р</i>	R, r
Б б	<i>Б б</i>	B, b	С с	<i>С с</i>	S, s
В в	<i>В в</i>	V, v	Т т	<i>Т т</i>	T, t
Г г	<i>Г г</i>	G, g	У у	<i>У у</i>	U, u
Д д	<i>Д д</i>	D, d	Ф ф	<i>Ф ф</i>	F, f
Е е	<i>Е е</i>	Ye, ye; E, e*	Х х	<i>Х х</i>	Kh, kh
Ж ж	<i>Ж ж</i>	Zh, zh	Ц ц	<i>Ц ц</i>	Ts, ts
З з	<i>З з</i>	Z, z	Ч ч	<i>Ч ч</i>	Ch, ch
И и	<i>И и</i>	I, i	Ш ш	<i>Ш ш</i>	Sh, sh
Й й	<i>Й й</i>	Y, y	Щ щ	<i>Щ щ</i>	Shch, shch
К к	<i>К к</i>	K, k	Ъ ъ	<i>Ъ ъ</i>	"
Л л	<i>Л л</i>	L, l	Ы ы	<i>Ы ы</i>	Y, y
М м	<i>М м</i>	M, m	Ь ь	<i>Ь ь</i>	'
Н н	<i>Н н</i>	N, n	Э э	<i>Э э</i>	E, e
О о	<i>О о</i>	O, o	Ю ю	<i>Ю ю</i>	Yu, yu
П п	<i>П п</i>	P, p	Я я	<i>Я я</i>	Ya, ya

*ye initially, after vowels, and after ъ, ь; e elsewhere.
When written as ë in Russian, transliterate as yë or ë.

RUSSIAN AND ENGLISH TRIGONOMETRIC FUNCTIONS

Russian	English	Russian	English	Russian	English
sin	sin	sh	sinh	arc sh	sinh ⁻¹
cos	cos	ch	cosh	arc ch	cosh ⁻¹
tg	tan	th	tanh	arc th	tanh ⁻¹
ctg	cot	cth	coth	arc cth	coth ⁻¹
sec	sec	sch	sech	arc sch	sech ⁻¹
cosec	csc	csch	csch	arc csch	csch ⁻¹

Russian English

rot curl
lg log

GRAPHICS DISCLAIMER

All figures, graphics, tables, equations, etc. merged into this translation were extracted from the best quality copy available.

Page 1.

FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC BAND.

V. M. Volkov.

Pages 2-4.

No typing.

Page 5.

Preface.

The rapid development of radio electronics and a deep penetration of it virtually into all areas of science and technology stipulated the need for the creation of amplifiers with the functional amplitude characteristic in the broad dynamic band of a change in the signal. The fields of application of functional amplifiers are very diverse: radar, ground-based and space radio communication, technique of measurement, automated control systems,

computers, etc. Without the functional amplifiers not at all conceived the execution of the device/equipment, which simulates living organism.

To questions of the expansion of the dynamic range of amplifiers, in particular, to the automatic gain control (ARU) are devoted many works. For example, in the latter/last five years to questions of ARU only in the transistorized amplifiers are devoted more than 200 Soviet and foreign works. However, in the majority of the cases the use/application of ARU provides for volume compression of output effect without taking into account the form of amplitude characteristic only.

At present in the technical literature from the functional amplifiers logarithmic amplifiers [7, 8] are sufficiently widely described. However, until now, are not presented general/common/total theory and design procedure of amplifiers with FAKh of any type in the broad dynamic band both on the tubes and on the transistors.

Page 6.

In this book is made the attempt to complete this gap/spacing and to give general/common/total approach to the design of functional

amplifiers independent of the type of amplifier instrument.

The material, published in the book, is the result of the 10-year work of the author and is to a considerable extent original.

Observations and wish about the book we request to direct to address: Kiev, 4, Puskinskaya, 28, publishing house.

Page 7.

Chapter 1.

METHODS OF OBTAINING THE FUNCTIONAL AMPLITUDE CHARACTERISTICS IN AMPLIFIERS.

§1. Criteria of evaluation of amplifiers with broad dynamic band.

The creation of functional amplifiers with the broad dynamic band (ShDD) on the input effect ν ¹ is at present urgent problem.

FOOTNOTE ¹. By input effect ν and output effect ϵ for the electronic amplifiers should be understood one of three values: voltage/stress, current or power in the dependence on the designation/purpose of functional amplifier. ENDFOOTNOTE.

This problem even more is complicated in the case of applying as the amplifier instruments the transistors, which have a small dynamic range on the input effect. In spite of that even number amplifiers with the broad dynamic band already extensively are used, up to now there is no general/common/total criterion, by using which it would be possible to consider different amplifiers with ShDD. By the

dynamic range of the input effect of amplifier, or dynamic range of amplifier, is understood the ratio of the maximum level of input effect v_{max} to minimum level v_{min} (Fig. 1):

$$d = \frac{v_{\text{max}}}{v_{\text{min}}} = \frac{v_{\text{max}}}{v_{\text{in}}}. \quad (1-1)$$

Value v_{max} corresponds to the level, on which the amplifier loses its amplifier properties. Value v_{min} is determined by the level, on which output effect e_{min} is sufficient for the normal work of the device/equipment, connected at the output of amplifier. Usually v_{min} is determined by the inherent noise level of amplifier, i.e., $v_{\text{min}} = v_{\text{in}}$.

Usual one- and multistage linear amplifiers will be considered. Let us assume that linear amplifier stage has amplitude characteristic, depicted in Fig. 1 (curve 1), and its dynamic range is determined by expression (1-1).

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Transmission factor of cascade/stage K.

Then for n-cascade amplifier the maximum input effect

$$E_{\text{max}} = \frac{v_{\text{max}}}{K^{n-1}};$$

the minimum input effect

$$E_{\text{min}} = v_{\text{in}} \gamma > v_{\text{in}}.$$

With $K \gg 1$ it is possible to take

$$E_{\text{max}} = v_{\text{m}} = v_{\text{max}},$$

then the dynamic range of n-cascade amplifier

$$D = \frac{E_{\text{max}}}{E_{\text{min}}} = \frac{v_{\text{max}}}{K^{n-1} v_{\text{min}}} = \frac{d}{K^{n-1}}, \quad (1-2)$$

i.e. K^{n-1} once is less than the dynamic range of cascade/stage.

Let us assume that for amplifier stage, carried out on the tube, $v_{\text{max}} = 5 \text{ V}$, $v_{\text{m}} = 5 \mu\text{V}$, $K=10$. Then according to equation (1-1) $d=100 \text{ dB}$. We accept a number of cascades/stages $n=5$. Then according to expression (1-2) $D_s=20 \text{ dB}$.

From the given examples it is evident that it is difficult to judge, what amplifier is better from the point of view of its dynamic and amplifier properties. For the quantitative estimation of the amplifier and dynamic properties of amplifiers, assembled on this type of amplifier instruments, it is expedient to introduce the concept of the dynamic quality:

for the cascade/stage

$$q_n = Kd; \quad (1-3)$$

for n-cascade amplifier

$$Q_n = K^n D. \quad (1-4)$$

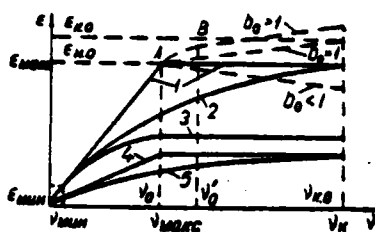


Fig. 1. Amplitude characteristics of linear and nonlinear amplifiers.

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Substituting in formulas (1-3), (1-4) values d and D , we obtain

$$q_A = Q_A,$$

i.e., with the linear amplification the quality one- and multistage amplifiers is identical. Now it is possible to claim that that linear amplifier is better, whose dynamic quality is more.

However, this confirmation it is insufficient for the functional (nonlinear) amplifiers. Let us show this based on the example of single-stage amplifier. Let us assume that functional amplifier stage consists of linear amplifier stage (U) and functional unit (FE), which can be connected in series on the diagram in Fig. 2a or Fig. 2b. Let us consider several simplest versions of the functional amplifiers:

1. Linear amplifier stage has following data: K' , d' , $q_A = K'd'$

(curve 1 in Fig. 1). The functional unit, which forms amplitude characteristic, is connected at the output of amplifier (Fig. 2a). In this case the effect v_{max} remained as before, since it is determined by amplifier, it is more accurate, by amplifier instrument (curve 3 in Fig. 1). For the functional amplifier we have:
 $K_{\phi y} = K'$; $d_{\phi y} = d'$; $q_{\text{L} \phi y} = q'_{\text{L}}$, i.e. dynamic range and dynamic quality was not changed in comparison with the linear amplifier.

2. Functional unit is connected at input of amplifier (Fig. 2b). In this case maximum value $v_{\text{H}} > v_{\text{max}}$ (curve 2 in Fig. 1).

Let us assume that $v_{\text{H}} = m v_{\text{max}}$. Then with the equality of initial transmission factors K we have: $q_{\text{L} \phi y} = m q'_{\text{L}} = m q'_{\text{L} \phi y}$, i.e. $q_{\text{L} \phi y} > q'_{\text{L} \phi y}$.

3. Amplifier has data: $K'' = K'/m$; v_{max} , v_{max} , $q_{\text{L}} = q'_{\text{L}}$ (curve 4 in Fig. 1). Functional amplifier is assembled on the block diagram Fig. 2b. It has following data: $K_{\phi y} = K'' = \frac{K'}{m}$; $v_{\text{max}} = v_{\text{H}} = m v_{\text{max}}$; $q_{\text{L} \phi y} = q'_{\text{L} \phi y}$.

As can be seen from the given examples, dynamic quality of FU, assembled on the identical amplifier instruments, depends on the structure of construction of FU, and with the identical structure of construction of FU its dynamic quality depends on the quality of linear amplifier stage.

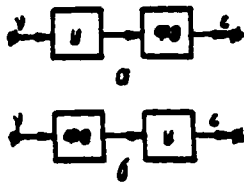


Fig. 2. Possible versions of the construction of the schematics of functional amplifiers.

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With the linear load the quality of amplifier stage can be replaced with the quality of amplifier instrument.

For a comparative evaluation of dynamic qualities of FU of identical structure it is expedient to introduce the concept of the standardized/normalized quality

$$H = \frac{Q_R}{q_R}, \quad (1-5)$$

which does not depend on the quality of amplifier instrument (linear cascade/stage). Then for the linear amplifier

$$H_{\text{лин}} = \frac{Q_{\text{л. лин}}}{q_{\text{л. лин}}} = 1.$$

Thus, by amplifier with the broad dynamic band should be understood amplifiers with the standardized/normalized dynamic quality of more than one, i.e., for them is satisfied the condition

$$H > 1. \quad (1-6)$$

In the determination of amplifier accepted with ShDD is not

considered the form of amplitude characteristic (AKh), since the expansion of dynamic range on the input effect only ensures the amplification of all possible levels of effect without the observance of specific ratios between the output effect and the input effect. By form of AKh with this input dynamic range is determined dynamic range on the output effect, which depending on type of FAKh can be less or more than dynamic range on the input effect.

According to the structure multistage FU can be consecutive or parallel types. Consecutive type functional amplifiers have high dynamic quality and therefore more greatly they are spread.

§2. Classification of functional amplifiers.

The functional amplitude characteristic of amplifier in general form can be registered as certain function $\epsilon = f(\nu)$ output effect of ϵ from the input effect ν .

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On the base of the achievements of contemporary radio electronics in the amplifiers it is possible to realize the broad class of functional dependences. From a mathematical point of view the functional dependences, which can be realized in the electronic

amplifiers, are divided into three classes: algebraic, transcendental and periodic.

By period should be understood the cycle of a change in the output effect ϵ with a change (increase) in the input effect ν in the dynamic range. In each period the function $\epsilon=f(\nu)$ can be described by algebraic or transcendental function.

Physical essence of FU is most fully reflected by this parameter as slope/transconductance FAKh, or, which is the same, by the differential transmission factor of amplifier, by the numerically equal to the ratio infinitesimal increment in the output effect of amplifier to infinitesimal increment in the input effect:

$$b = \frac{\partial f(\nu)}{\partial \nu} = \frac{\partial \epsilon}{\partial \nu}. \quad (1-7)$$

According to the character of coefficient b it is possible to judge the possible methods of realization of FAKh. Therefore it is expedient to additionally class FU on the base of differential transmission factor. Outcome from this principle, entire variety of FU can be divided into the following fundamental types:

1. Amplifiers with the constant differential transmission factor over the dynamic range

$$b = \text{const.} \quad (1-8)$$

2. Amplifiers with variable/alternating/variable and reduced b

$$b = \text{var}(\downarrow). \quad (1-9)$$

over dynamic s-band increase in input effect ν

3. Amplifiers with variable/alternating/variable and increasing b with increase/growth ν

$$b = \text{var}(\uparrow). \quad (1-10)$$

Mathematically differential transmission factor of FU over the dynamic range can be described by algebraic or transcendental function. In periodic FAKh one of conditions (1-8), (1-9) or (1-10) is realized.

The first type of FU includes the amplifiers with the linear amplitude characteristic (lin AKh)

$$f(\nu) = K_{\text{av}}. \quad (1-11)$$

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To the second type of FU can be attributed the following special cases:

a) logarithmic amplifiers (LAKh)

$$f(\nu) = c \ln \gamma \nu; \quad (1-12)$$

b) amplifiers with the exponential amplitude characteristic (SAKh) with $\beta < 1$ (SAKh)

$$f(\nu) = a \nu^{\beta} \quad (1-13)$$

and so forth.

The third type of FU includes:

a) amplifiers with the exponential (exponential) amplitude characteristic (PAKh, EAKh)

$$f(v) = aM^{\beta} \text{ or } f(v) = ae^{\beta v}; (1-14)$$

b) amplifiers with the exponential amplitude characteristic with $\beta > 1$ (SAKh,) and so forth.

Any real FU under the sufficiently small input influence works in the linear conditions, and then with some fully specific level of input effect v_* passes into the nonlinear operating mode. If level v_* exceeds the level of its own input noises, then FU has linearly-functional amplitude characteristic.

Some special cases of FAKh of real amplifiers are shown in Fig. 3a, and 3b a change in the differential transmission factors different FU with a change in the input effect is shown.

Transition from the linear section of characteristic to the nonlinear occurs monotonically or with the fracture (at angle)

depending on value and character of a change in the differential transmission factor with the work of FU in the nonlinear mode/conditions. The most pronounced fractures (angles) must have characteristics of the amplifier-limiters (Fig. 3a), for which must be implemented equalities $b_{\text{max}} = \text{const}$, $b_{\text{max}} = b_{\text{orp}} = 0$ and multilevel (two-, three- and n-level) amplifiers with linearly stepped characteristic (Fig. 4).

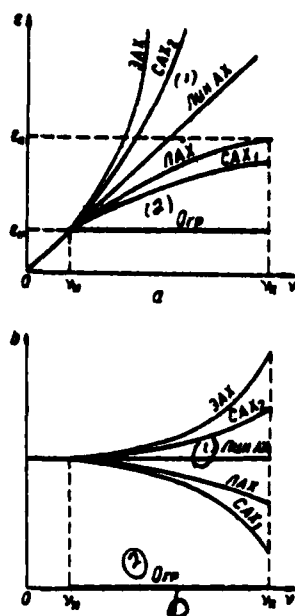


Fig. 3. Characteristics of the functional amplifiers: a) amplitude; b) differential.

Key: (1). Lin. (2). Limit.

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For periodic FAKh mutual conductance periodically is changed (it grows or decreases) according to the specific law with an increase in the input effect ν . Transition from one law of change $\delta = \text{var}(\uparrow)$ to the next $\delta = \text{var}(\uparrow)$ or $\delta = \text{const}$ can occur smoothly (case of monotone characteristics), with the fracture it is abrupt (with the disruption in the characteristic). In practice discontinuous FAKh more

frequently are applied, when coefficient b , being changed according to the specific law in some assigned range of a change in the input effect ν , after achieving determined of value b_{\max} when $b = \text{var}(\uparrow)$ or b_{\min} when $b = \text{var}(\downarrow)$, abruptly it is changed, being returned to its initial value, or some new value accepts, it is more or less than the initial (Fig. 5).

Let us agree to call the characteristics, which are a special case of a broader class of periodic FAKh, functional-discontinuous.

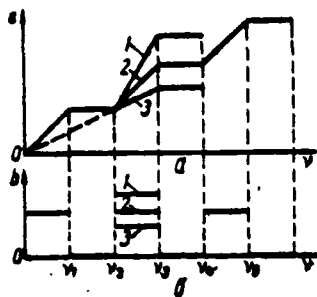


Fig. 4.

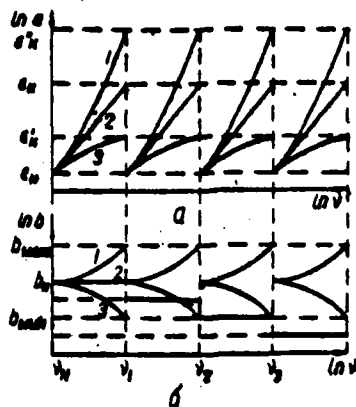


Fig. 5.

Fig. 4. Polygonal characteristics: a) amplitude; b) differential; 1 - $b_1 \gg b_2$; 2 - $b_1 = b_2 = b_3$; 3 - $b_1 < b_2$.

Fig. 5. Periodic (discontinuous) characteristics: a) amplitude; b) differential; 1 - exponential; 2 - linear; 3 - logarithmic.

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If coefficient b abruptly is changed at the values $\nu = \nu_1, \nu_2, \dots, \nu_n$, then we will call the relations

$$\frac{\nu_1}{\nu_2} = d_1, \frac{\nu_2}{\nu_1} = d_2, \dots, \frac{\nu_n}{\nu_{n-1}} = d_n \quad (1-15)$$

periods or dynamic ranges of change ν , which fall for one adjustment. For simplification in the construction/design of FU usually they choose

$$d_1 = d_2 = \dots = d_{n-1} = d_n = d. \quad (1-16)$$

If the depth of the adjustment of coefficient of b

$$p_i = \frac{b_{\max i}}{b_n}, \text{ or } p_i = \frac{b_n}{b_{\min i}}, \quad (1-17)$$

the change in coefficient of b for the amplifier with the periodic (by discontinuous FAKh) analytically it is possible to write:

with $b(v) = \text{var}(\uparrow)$

$$b_n(\uparrow) = \frac{b(v)}{\prod_{i=1}^m p_i}; \quad (1-18)$$

with $b(v) = \text{var}(\downarrow)$

$$b_n(\downarrow) = b(v) \prod_{i=1}^m p_i, \quad (1-19)$$

where $i = \alpha \lg \frac{v}{v_n} = 1, 2, \dots, m$ - whole number; α - coefficient of standardization. In the particular case with $d=10$ the coefficient $\alpha=1$;

with $b(v) = \text{const}$ coefficient b can be changed according to the law (1-18) or (1-19), i.e., abruptly be reduced or increase.

If the coefficients of adjustment p_i are equal to each other, i.e.

$$p_1 = p_2 = \dots = p_i = p_n = p,$$

then expressions (1-18) and (1-19) take the form

$$\begin{aligned} b_n(\uparrow) &= \frac{b(\nu)}{p^k}; \\ b_n(\downarrow) &= b(\nu) p^l. \end{aligned}$$

Let us find the general/common/total expression for FAKh of real amplifier under the assigned law of a change in coefficient of $b(\nu)$ in the dynamic range. With $\nu < \nu_n$ FU works in the linear conditions with the initial transmission factor K_n . In this case $b(\nu) = K_n$ and amplitude characteristic is described by the expression

$$\begin{aligned} f(\nu) &= \nu K_n. \\ 0 &< \nu < \nu_n. \end{aligned} \quad (1-20)$$

With $\nu = \nu_n$ FU passes into the functional operating mode.

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So that the amplitude characteristic of amplifier would have the smooth transition (without the sharp fractures), at any transition point of it of two adjacent sections must be implemented the following conditions:

- a) the equality of the ordinates of the adjacent sections

$$\lim_{\Delta\nu \rightarrow 0} f(\nu - \Delta\nu) = \lim_{\Delta\nu \rightarrow 0} f(\nu + \Delta\nu); \quad (1-21)$$

- b) the equality of first-order derivatives, or the equality of the differential transmission factors infinitesimal adjacent sections

$$\frac{\partial f(\nu - \Delta\nu)}{\partial \nu_{\Delta\nu \rightarrow 0}} = \frac{\partial f(\nu + \Delta\nu)}{\partial \nu_{\Delta\nu \rightarrow 0}}; \quad (1-22)$$

- c) the identical signs of first-order derivatives.

With the work of amplifier in the functional mode/conditions of AKh it is written/recorded in the form

$$f(v) = \int b(v) dv + C = \phi(v) + C. \quad (1-23)$$

After finding constant of integration C, on the basis of initial conditions (1-21) and (1-22) when $v = v_n$, finally we obtain

$$f(v) = f(v_n) + \Delta\phi(v), \quad (1-24)$$

where

$$\Delta\phi(v) = \phi(v) - \phi(v_n), \quad \phi(v) = \int b(v) dv.$$

Expression (1-24) is general/common/total and it is useful for finding the mathematical expression of FAKh of real amplifier according to the assigned function of a change in the differential transmission factor of FU.

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§3. Qualitative indices of functional amplifiers.

Fundamental qualitative indices of FU are: dynamic range of FAKh on the input effect

$$d' = \frac{v_n}{v_n}, \quad (1-25)$$

where v_n and v_n - input effects, with which it begins and is finished by FAKh of the amplifier:

dynamic range of FAKh on the output effect

$$d_{\text{max}} = \frac{f(v_n)}{f(v_m)} = \frac{\epsilon_n}{\epsilon_m}, \quad (1-26)$$

where ϵ_n and ϵ_m - value of output effect respectively at the values of input effect v_n and v_m ;

contraction coefficient C when $b = \text{var}(\downarrow)$ or expansion P when $b = \text{var}(\uparrow)$

slope/transconductance of FAKh or differential transmission factor FU

$$b = \frac{\partial \epsilon}{\partial v};$$

slope/transconductance of FAKh in the converted coordinates ²

$$a = \frac{\Delta[\epsilon]}{\Delta[v]}, \quad (1-28)$$

where $\Delta[\epsilon]$ and $\Delta[v]$ value of ϵ and v in the converted coordinates.

FOOTNOTE ¹. Let us agree to designate the dynamic range of FAKh of multistage amplifier through D.

². Examples of transformation of coordinates will be given in examination of concrete/specific/actual types FAKh. ENDFOOTNOTE.

In practice linear coordinates ϵ and v are converted in such a

way that the condition $\sigma = \text{const}$ with the change ν would be satisfied;

the accuracy of realization of FAKh:

a) in the absolute divergence of objective parameter $f_p(\nu)$ from precise $f_r(\nu)$

$$\Delta f = \Delta f(\nu_i) = f_r(\nu_i) - f_p(\nu_i); \quad (1-29)$$

b) according to the relative deflection

$$\delta f(\nu_i) = \frac{\Delta f}{f_i}; \quad (1-29a)$$

c) according to the absolute divergence of slope/transconductance of FAKh

$$\Delta \sigma(\nu_i) = \sigma_r(\nu_i) - \sigma_p(\nu_i); \quad (1-29b)$$

d) according to the relative deflection of the slope/transconductance

$$\delta \sigma(\nu_i) = \frac{\Delta \sigma(\nu_i)}{\sigma_r(\nu_i)}; \quad (1-29b)$$

the initial transmission factor of amplifier with the work in linear conditions K_u ;

resonance frequency f_0 and the passband Π of selective amplifiers, upper f_{max} and lower f_{min} cut-off frequencies for the aperiodic amplifiers in linear mode/conditions;

dynamic quality factor and the standardized/normalized dynamic quality, determined respectively by expressions (1-3)-(1-5);

stability and the recurrence of FAKh, which characterizes the possibility of serial production of FU with the identical parameters.

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After using formulas (1-24)-(1-29), let us find expressions and qualitative indices for the most widely used types of FAKh. For obtaining the generalized results let us introduce the standardized values of the input effect x and the output effect z :

$$z = \frac{v}{v_n}, \quad (1-30)$$

$$x = \frac{e}{e_n} = \frac{e}{v_n K_n}. \quad (1-31)$$

Expressions for the differential transmission factor $b(v)$, the usual and standardized/normalized amplitude characteristics are given in Table 1.

Linear amplifier.

For the linear amplifier is implemented the equality

$$D_{\text{out}} = D. \quad (1-32)$$

The lower level of input effect in the linear amplifier is limited to internally-produced noise. Therefore in the case of FU of voltage/stress for minimum input effect v_n they accept the

voltage/stress of internally-produced noise, converted to the input of amplifier, i.e.

$$v_n = U_{n, n} = U_{n, n}.$$

The upper level of input effect v_n is limited to the linear section of the passage characteristic of amplifier instrument.

The dynamic range of linear amplifier can be widened due to a decrease of initial v_n and an increase in final v_n of input effects with a simultaneous increase in final output effect v_n , since condition (1-32) is satisfied.

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Table 1.

(2) Амплитудная характеристика	(3) Дифференциальный коэффициент передачи $\frac{du}{dV}$	(1) Имя цепи		(7) Обратная зависимость $s = f(u)$	(8) Примечание
		(4) Математическая запись ФАХ реального усилителя	(6) Нормированная $u = f(x)$		
(9) Линейная $u = bV$	K_u	$u = K_u V$	$u = x$	$x = u$	$u = \frac{V}{V_u}; \frac{u}{u_u} = \frac{V}{V_u}; u_u = V_u K_u$
(10) Логарифмическая $u = a_N \ln(V)$	$\frac{a_N}{V} = \frac{a K_u}{x}$	$u = K_u V \left(a \ln \frac{V}{V_u} + 1 \right)$	$u = a \ln x + 1$	$b = \exp \frac{u-1}{a}$	$u = u_u; a_N = a u_u;$ $a = \frac{1}{\ln N}; \gamma = \frac{N}{V}$
(11) Показательная $u = a M^V$	$K_u M^{V-1}$	$u = K_u V_u \left[1 + \frac{1}{V_u \ln M} (M^V - 1) \right]$	$u = 1 + \frac{1}{V_u \ln M} (M^x - 1)$	—	$u = \frac{p}{V_u}$
(12) Экспоненциальная $u = a e^{bV}$	$K_u \exp \left[\frac{1}{a} \left(\frac{V}{V_u} - 1 \right) \right]$	$u = K_u V_u \exp \left[\frac{1}{a} \left(\frac{V}{V_u} - 1 \right) \right]$	$u = \exp \frac{x-1}{a}$	$x = a \ln u + 1$	$u_u = \beta = \frac{\ln N}{V_u}; a = \frac{K_u V_u}{N}$
(13) Степенная $u = a V^\beta$	$\beta = \frac{m}{n}$	$K_u \sqrt[n]{\left(\frac{V}{V_u} \right)^{m-n}}$	$u = K_u V_u \left[\frac{n}{m} \sqrt[n]{\left(\frac{V}{V_u} \right)^m} - \frac{n}{m} + 1 \right]$	$u = \frac{n}{m} \sqrt[n]{x^m} - \frac{n}{m} + 1$	$x = \sqrt[n]{\left(\frac{m}{n} u - \frac{m-n}{n} \right)^n}$
	$\beta = \frac{1}{n}$	$K_u \sqrt[n]{\left(\frac{V}{V_u} \right)^{1-n}}$	$u = K_u V_u \left(n \sqrt[n]{\frac{V}{V_u}} - n + 1 \right)$	$u = n \sqrt[n]{x} - n + 1$	$x = \left(\frac{u-1+n}{n} \right)^n$
	$\beta = m$	$K_u \left(\frac{V}{V_u} \right)^{m-1}$	$u = K_u V_u \left[\left(\frac{V}{V_u} \right)^m \frac{1}{m} - \frac{1}{m} + 1 \right]$	$u = \frac{x^m + m - 1}{m}$	$x = \sqrt[m]{m u - m + 1}$
(14) Усилитель-ограничитель	(15) K_u при $V < V_u$ b_0 при $V > V_u$	$u = \begin{cases} K_u V & \text{при } V < V_u \\ K_u V_u + b_0 (V - V_u) & \text{при } V > V_u \end{cases}$	$u = \begin{cases} x & \text{при } x < 1 \\ a_0 + b_0 (x - 1) & \text{при } x > 1 \end{cases}$		$u = \frac{V}{V_u}; u = \frac{1}{a_0}$
(16) Линейно-прерывистая	$\frac{K_u}{\prod_{i=1}^n \rho_i}$	$u = V \frac{K_u}{\prod_{i=1}^n \rho_i}$	$\frac{x}{\prod_{i=1}^n \rho_i}$	$u = x$	$u = a \lg a = 1, 2, 3, \dots$
(17) Логарифмо-прерывистая	$\frac{a K_u V_u}{V} \prod_{i=1}^n \rho_i$	$u = K_u V_u \left(a \ln \frac{V}{V_u \prod_{i=1}^n \rho_i} + 1 \right)$	$a \ln \frac{x}{\prod_{i=1}^n \rho_i} + 1$	$x = \exp \left(\frac{u-1}{a} \right) \prod_{i=1}^n \rho_i$	

Key: (1). Parameter. (2). Amplitude characteristic. (3). Differential transmission factor. (4). Mathematical recording of FAKh of real amplifier. (5). complete. (6). calibrated. (7). Inverse dependence. (8). Note. (9). Linear. (10). Logarithmic. (11). Exponential. (12). Exponential. (13). Exponential. (14). Amplifier-limiter. (15). with.

(16). linear-discontinuous. (17). logarithm-discontinuous.

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Therefore for expanding the range D of linear amplifier it is necessary: to apply the low-noise amplifier instruments (UP); to apply UP with the large linear section of passage characteristic; to compensate the nonlinearity of passage characteristic of UP; to apply the negative feedback, which linearizes passage characteristic of UP; to repeatedly use a linear section of passage characteristic of UP.

The logarithmic amplifier

During the logarithmic operation of signal with the Napierian base e the differential amplifier gain is changed according to the law

$$b_e(v) = \frac{\partial e}{\partial v} = \frac{e}{v} = \frac{K_E}{s} \quad (1-33)$$

and amplitude characteristic according to equation (1-19) is described by the expression

$$\begin{aligned} f(v) = e &= f(v_E) + \Delta\phi(v) = \\ &= K_E v_E (\ln x + 1). \end{aligned} \quad (1-34)$$

In this case the standardized/normalized characteristic

$$z = \ln x + 1. \quad (1-35)$$

If we along the axis of abscissas plot values v (or x) on the logarithmic scale, and along the axis of ordinates - value e (or z)

on the graphic scale, then the curves, described by expressions (1-34) and (1-35), will be depicted in the form of straight lines (segment AB in Fig. 6). Let us agree to call this converted scale 1st type semilogarithmic scale. Logarithmic amplitude characteristic on the semilogarithmic scale has constant slope/transconductance σ . If we straight/direct AV continue to the side of smaller voltages/stresses, then it will intersect the axis of abscissas at point ν_0 . ν_0 physically corresponds to the input effect, under which at the output of ideal logarithmic amplifier output effect is equal to zero. It is easy to find value ν_0 , if the right side of expression (1-34) is made equal to zero:

$$\nu_0 = \frac{\nu_{\pi}}{\sigma}. \quad (1-36)$$

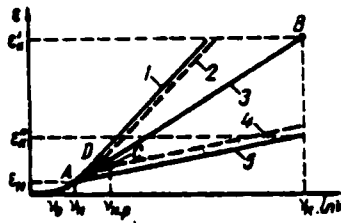


Fig. 6. The logarithmic characteristics of real amplifier on 1st type semilogarithmic scale: 1, 3, 5 - precise; 2, 3, 4 - real; 1 - $N < e$ ($a > 1$); 3 - $N = e$ ($a = 1$); 5 - $N > e$ ($a < 1$).

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Slope/transconductance σ of the logarithmic characteristic, depicted on semilogarithmic scale,

$$\sigma = \frac{\Delta(e)}{\Delta(u)} = \frac{\Delta e}{\ln D} = \frac{e_2 - e_1}{\ln D}. \quad (1-37)$$

In the case of FU voltage after substitution

$$e_2 = U_{\text{max.}} = U_{\text{max.}} (\ln D + 1)$$

$$\sigma = U_{\text{max.}} \left[\frac{e^{\ln D}}{\ln D} \right]. \quad (1-38)$$

Key: (1) V/neper.

Slope/transconductance σ can be calculated also in volts to decibels [V/dB]. There is a following dependence between the values of slope/transconductance, expressed in different units,

$$\sigma \left[\frac{(1)}{\text{dB}} \right] = \frac{\sigma \left[\frac{(2)}{\text{neper}} \right]}{8.68}. \quad (1-39)$$

Key: (1). [V/dB]. (2). [V/neper].

The logarithmic amplitude characteristic of amplifier is ideally precise, if in entire dynamic range D condition $\sigma = \text{const}$ is satisfied.

From the given analysis it is evident that in the case of logarithmic operation according to the law of natural logarithm with foundation $N=e=2.72$ the slope/transconductance of LAKh on the semilogarithmic scale is numerically equal to output effect ϵ_n , with which it begins with LAKh. Consequently, changing value ϵ_n , it is possible to vary slope/transconductance of LAKh. Since is implemented the equality

$$\epsilon_n = K_n v_n = \epsilon_n$$

then at given values ϵ_n and K_n LAKh of amplifier must begin at the completely specific value of the input effect

$$v_n = \frac{\epsilon_n}{K_n}$$

and vice versa, at given values ϵ_n and v_n the amplifier gain with the work in the linear conditions must be the completely specific value

$$K_n = \frac{\epsilon_n}{v_n}$$

The characteristics of logarithmic amplifier in the semilogarithmic scale for three values of slope/transconductance σ at constant value v_n and the different values of factor of amplification K_n are given in Fig. 7a; at the constant value of coefficient K_n and different values v_n - in Fig. 7b.

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From the graphs it is evident that in all cases the linear section of characteristic smoothly passes into the logarithmic. Dynamic range in the output effect

$$D_{\text{max}} = \ln D + 1; \quad (1-40)$$

contraction coefficient of the signal having taken the logarithm of

$$C_s = \frac{D}{D_{\text{max}}} = \frac{D}{\ln D + 1}. \quad (1-41)$$

From expressions (1-40) and (1-41) we see that for the given value of dynamic range on the input effect D values D_{max} and C_s for the characteristics with the different value of slope/transconductance and foundation e are constant values. This is a deficiency/lack in the characteristic, since in many instances it is necessary at the given value of D to have different values D_{max} . The characteristic, with which input signal in the amplifier takes the logarithm of with any logarithm to the base N , possesses this property.

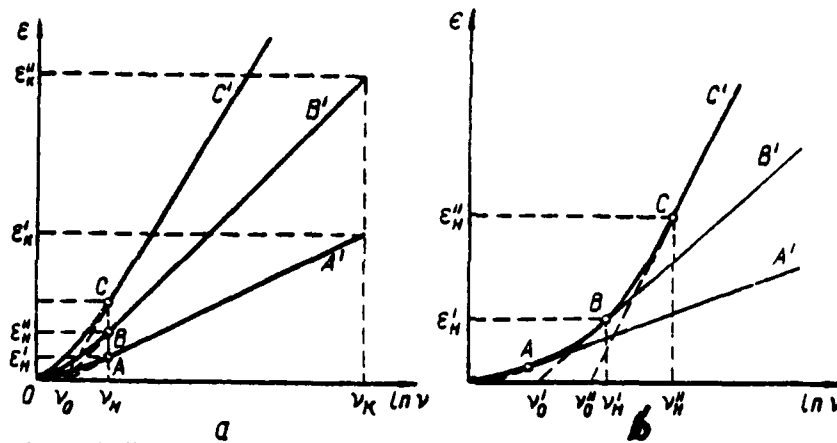


Fig. 7. Logarithmic characteristics of amplifier at the different values of the slope/transconductance: a) $v_H = \text{const}$; $K_H = \text{var}$; b) $K_H = \text{const}$; $v_H = \text{var}$.

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Mathematically this characteristic can be registered as follows:

$$\varepsilon = \varepsilon_N + \sigma_N \ln \frac{v}{v_N} = \varepsilon_N (a \ln x + 1), \quad (1-42)$$

where $\sigma_N = a\varepsilon_N$ - slope/transconductance of LAKh with any foundation N;
 $a = 1/\ln N$ - conversion factor from foundation e of natural logarithm to any foundation.

With $N=e$ coefficient $a=1$.

Let us agree this characteristic to call, the generalized logarithmic amplitude characteristic, also, near the designations of the parameters of amplifier to place index N.

Standardized/normalized generalized LAKh

$$z = a \ln z + 1. \quad (1-43)$$

Dynamic range in output effect $D_{N_{\text{max}}}$ and the contraction coefficient of signal

$$D_{N_{\text{max}}} = a \ln D + 1; \quad (1-44)$$

$$C_N = \frac{D}{a \ln D + 1}. \quad (1-45)$$

Thus, changing the value of the logarithm to the base N of characteristic, it is possible to change value of D and C . With increase in N (decrease a) D_N is reduced, and C_N increases. This is clearly illustrated by Fig. 6, on which are depicted three characteristics: with $N < e$ (1), $N = 2$ (2) and $N > e$ (3).

It is necessary to note that with $N \neq e (a \neq 1)$ the characteristic has a fracture at the point of the joint (Fig. 6, point A) of linear and logarithmic sections (curves 1 and 5 in Fig. 6). This is explained by the fact that with $N \neq e$ at the point of joint condition (1-22) is not satisfied. Theoretically this characteristic can be constructed. In practice characteristic with the sharp fracture in the real amplifier cannot be obtained, since the real nonlinear elements/cells can only smoothly change amplification factor. As a result this of nonlinear characteristic originates the

law, different from the logarithmic, but the ensuring smooth transition from the linear to the logarithmic (transition sections AD on curved 2 and AC in curved 4, Fig. 6).

In the presence of transition section working section of LAKh is shortened and is shifted/sheared into the region of large input effects. The extent of transition section can be different and is determined by the properties of the nonlinear elements/cells, used in the amplifier. With this type of nonlinear elements/cells the extent of transition section is more at the larger value of N.

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This must be had in mind during the development of logarithmic amplifiers and applied such nonlinear elements/cells, with which it is possible to obtain the smallest transition section of characteristic. It is obvious that the sharpest fracture in the amplitude characteristic can be obtained, applying nonlinear elements/cells with the large slope/transconductance of the initial section of volt-ampere characteristic (with small voltages/stresses on the nonlinear element/cell).

If at the disposal of designer is a logarithmic amplifier with the specific parameters, which do not satisfy technical

specifications, then some parameters can be changed, if to include/connect at input and output of logarithmic amplifier linear devices/equipment with the transmission factor of more than one (amplifiers) or less than one (attenuators). Upon the start of linear device/equipment at the output of logarithmic device/equipment with the transmission factor K' , the slope/transconductance of LAKh with the retention/preservation/maintaining of dynamic range on output effect D_{max} is changed. In this case the slope/transconductance of LAKh and output effects ϵ_1 and ϵ_n of total amplifier circuit are equal to

$$\sigma' = \sigma K'_0; \quad \epsilon'_1 = K'_0 \epsilon_1; \quad \epsilon'_n = \epsilon_n K'_0.$$

Working section of LAKh it is possible to move over the range of input effect without a change in values D , D_{max} and σ , including before the logarithmic amplifier linear device/equipment with the transmission factor K'' . In this case the changed parameters of LAKh of the total amplifier circuit

$$\nu'_n = \frac{\nu_n}{K''_0}; \quad \nu''_n = \frac{\nu_n}{K''_0}.$$

Amplifier with the exponential amplitude characteristic (PAKh and EAKh).

Expressions for the exponential amplitude characteristic and its parameters are given in Table 1. Let us consider one of them:

$$s = K_n v_n \left[1 + \frac{1}{v_n \beta \ln M} (M^{\beta(v-v_n)} - 1) \right]. \quad (1-46)$$

If we on the axis of abscissas plot graphic scale, and on the axis of ordinates - logarithmic, then dependence (1-46) when $v > v_n$ will be depicted in the form of straight line (Fig. 8).

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Let us agree this coordinate system to call 2nd type linear-logarithmic, and scale by 2nd type linear-logarithmic, or semilogarithmic scale.

Dependences (1.46) for different values of M and β are given in Fig. 8, from which it is evident that with $M > e$ (curve 2) and $M < e$ (curve 4) between the the linear OA and by the nonlinear sections (CC' and DD') are transition nonlinear sections AC and AD with the law, different from the exponential. Nonlinear sections with the linear ones are smoothly joined by these sections. The same is observed with $M = e$ and $\beta = \frac{1}{v_n}$ or $\beta < \frac{1}{v_n}$. Transition sections shorten the extent of the working section of exponential characteristic.

Of this deficiency/lack is deprived exponential dependence with $M = e$ and $\beta = \frac{1}{v_n}$ (curve 3 in Fig. 8)

$$s = K_n v_n \exp\left(\frac{v}{v_n} + 1\right). \quad (1-47)$$

If segment AB of dependence (1-47) is continued into the region of smaller input effects, then with $\nu=0$ it will intersect the axis of ordinates at point $\epsilon = \epsilon_0 = \frac{u_0}{\nu}$. Slope/transconductance of EAKh, depicted in the converted coordinates,

$$c_{\text{ex}} = \frac{\ln D_{\text{ex}}}{\Delta \nu}; \quad (1-48)$$

with $\nu=U$

$$c_{\text{ex}} = U_{\text{ex}} \left[\frac{u_0}{\epsilon} \right].$$

Thus, with foundation ϵ slope/transconductance of EAKh can be changed, varying the value of input effect ν_{ex} , at which it begins.

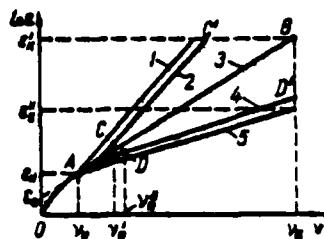


Fig. 8. The exponential characteristics of real amplifier in scale 2nd type semilogarithmic: 1 - $M=e$; $a>1$; 2 - $M>e$, $\beta+\frac{1}{v_0}$; 3 - $M=e$, $\beta-\frac{1}{v_0}$, $a=1$; 4 - $M<e$, $\beta+\frac{1}{v_0}$; 5 - $M=e$, $a<1$.

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Since $e_u = K_u v_u$, it is possible to consider two cases a change in the slope/transconductance σ :

$K_u = \text{const.}$ For increase e_u it is necessary to reduce value v_u and to simultaneously increase the transmission factor of amplifier K_u with the work in the linear conditions.

$\bar{K}_u = \text{const.}$ For increase e_u it is also necessary to reduce v_u . In this case e_u is reduced

This deficiency/lack does not have EAKh with any logarithm to the base N , described by the following expression:

$$i = i_s \exp\left(\frac{x-1}{a}\right), \quad (1-49)$$

where

$$a = \frac{1}{\ln N}.$$

In this case slope/transconductance EAKh

$$g_{av} = \frac{1}{av_u} = \frac{\ln N}{v_u}.$$

When $v_u = U_{sx, u}$

$$g_{av} = \frac{\ln N}{U_{sx, u}} \left[\frac{m\alpha n}{e} \right]. \quad (1-50)$$

The characteristics, described by expression (1-49), with $N \neq e$ have a fracture at the point of joint A of linear and nonlinear sections (curves 1 and 5 in Fig. 8). It is natural that real EAKh will have certain transition section and differ somewhat from the calculated ones. However, selecting the appropriate nonlinear elements/cells, it is possible to considerably shorten transition section and objective parameter to drive on sufficiently closely to calculated EAKh.

Amplifiers with EAKh, described by expression (1-49), most extensively are used, since in them it is possible to obtain the different value of slope/transconductance of EAKh at given values K_u , v_u and D. This fact, in turn, facilitates in practice the

necessary mating of amplifiers with LAKh and EAKh for obtaining the required parameters of electronic device as a whole (receivers RLS, which orient systems).

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Amplifiers with the exponential amplitude characteristic.

Ideal exponential amplitude characteristic is described by the expression

$$z = f(x) = \alpha x^\beta = \alpha x^{\frac{m}{n}}, \quad (1-51)$$

which with $\alpha=1$, $\beta>0$ and $x>0$ is represented as the graphs, given in Fig. 9a. If we switch over to logarithmic coordinates $\eta=\lg z$, $\xi=\lg x$, we will obtain the dependence

$$\eta = \beta \xi. \quad (1-52)$$

To the basis of expression (1-52) and Fig. 9b it is possible to introduce the following concept of the slope/transconductance of SAKh, depicted in the logarithmic system of coordinates

$$c_0 = \frac{\eta_2 - \eta_1}{\xi_2 - \xi_1} = \frac{\lg z_2 - \lg z_1}{\lg x_2 - \lg x_1} = \frac{\lg D_{\text{max}}}{\lg D} = \beta. \quad (1-53)$$

Expressions for the differential coefficient AKh are given in Table 1.

With $\beta < 1$ one should speak about the contraction coefficient of the reinforced signal C, while with $\beta > 1$ - about the coefficient of expansion P.

Amplifier-limiter.

The amplitude characteristic of ideal amplifier-limiter (UO) is depicted in Fig. 1 and it is mathematically described as follows:

$$s = f(v) = \begin{cases} K_{\text{н}} v & \text{при } v < v_0; \\ K_{\text{н}} v_0 & \text{при } v > v_0. \end{cases} \quad (1-54)$$

Key: (1). with.

where v_0 - input effect, under which the limitation begins.

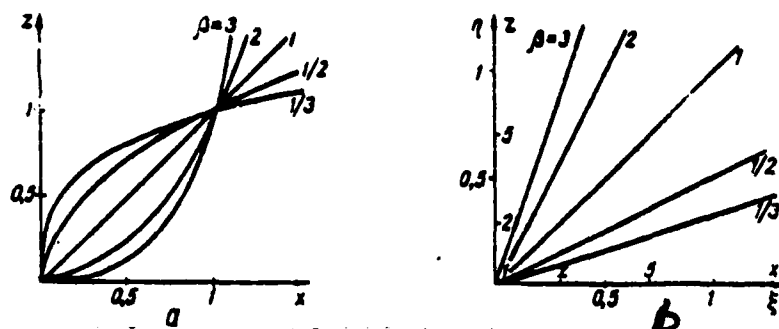


Fig. 9. Power characteristics at the different values of index β : a) graphic scale; b) logarithmic scale.

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The differential transmission factor of ideal amplifier-limiter must be equal to amplification factor with the work of amplifier in the mode/conditions of limitation, i.e.,

$$b_{yo} = \begin{cases} K_{\Sigma} = \text{const} & \text{при } v < v_0; \\ 0 = \text{const} & \text{при } v > v_0. \end{cases} \quad (1-55)$$

Key: (1). with.

In real UO differential transmission factor b , with the work of UO in the mode/conditions of limitation differs from zero and can be the value of positive and negative.

In this case amplitude characteristic of UO takes the form,

shown by prime in Fig. 1, and is described by the expression

$$\epsilon = \begin{cases} K_{\pi} v & \text{при } v \leq v_0; \\ K_{\pi} v_0 + b_0(v - v_0) & \text{при } v > v_0. \end{cases} \quad (1-56)$$

Key: (1). with.

Expressions for the standardized/normalized amplitude characteristic of real UO are given in Table 1.

Dynamic range of UO on the output effect

$$D_{o, \text{max}} = \frac{\epsilon_{\pi, o} - \epsilon_{\pi, o}}{\epsilon_{\pi, o}} = \frac{b_0(D_0 - 1)}{K_{\pi}}, \quad (1-57)$$

where $D_0 \frac{v_{\pi, o}}{v_0} = x_{\pi}$ - dynamic range of limitation on the input effect.

If according to the technical specifications are assigned D_0 , K_{π} and $D_{o, \text{max}}$, then differential coefficient with the work of UO in the mode/conditions of limitation must not exceed the value

$$b_0 < b_{0, \text{max}} = \frac{K_{\pi} D_{o, \text{max}}}{D_0 - 1}. \quad (1-58)$$

In the real amplitude characteristic of UO between the linear section and the section of limitation is a nonlinear section of the smooth transition (section AB in Fig. 1), which moves aside the threshold of limitation to the side of large input effects v_0 and reduces the dynamic range of limitation on input effect D_0 .

During the development of UO it is necessary to attempt to decrease the extent of transient section and to satisfy the condition $b_0 \rightarrow 0$.

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Amplifiers with periodic (discontinuous) FAKh.

On the basis of expressions (1-16) and (1-17) it is possible to do the conclusion that for obtaining the analytical expression for any discontinuous FAKh and its parameters it is necessary in the expressions, which describe FAKh and their parameters, value ν and x to divide into product $\prod_{i=1}^m p_i$, where $i = \alpha \lg x = 1, 2, \dots, m$.

With the gap count FU (adjustments of amplifier) m working section of FU is used $m+1$ times (Fig. 5). In this case the dynamic ranges Pr FAKh on the input effect and the output effect are respectively equal to

$$D = d^{m+1} = d^m,$$

or

$$D_{os} = (m+1) 20 \lg d; \quad (1-59)$$

$$D_{max} = d_{max}, \quad (1-60)$$

where d and d_{max} - dynamic ranges on input effect and output effect, that fall for one adjustment of amplifier (between two disruptions of

characteristic).

Contraction coefficient of the reinforced signal

$$C = \frac{D_{\max}}{D} = \frac{d_{\max}}{d}. \quad (1-61)$$

For linear-discontinuous characteristic $d_{\max} = d$. Then

$$C_{\max} = 1. \quad (1-62)$$

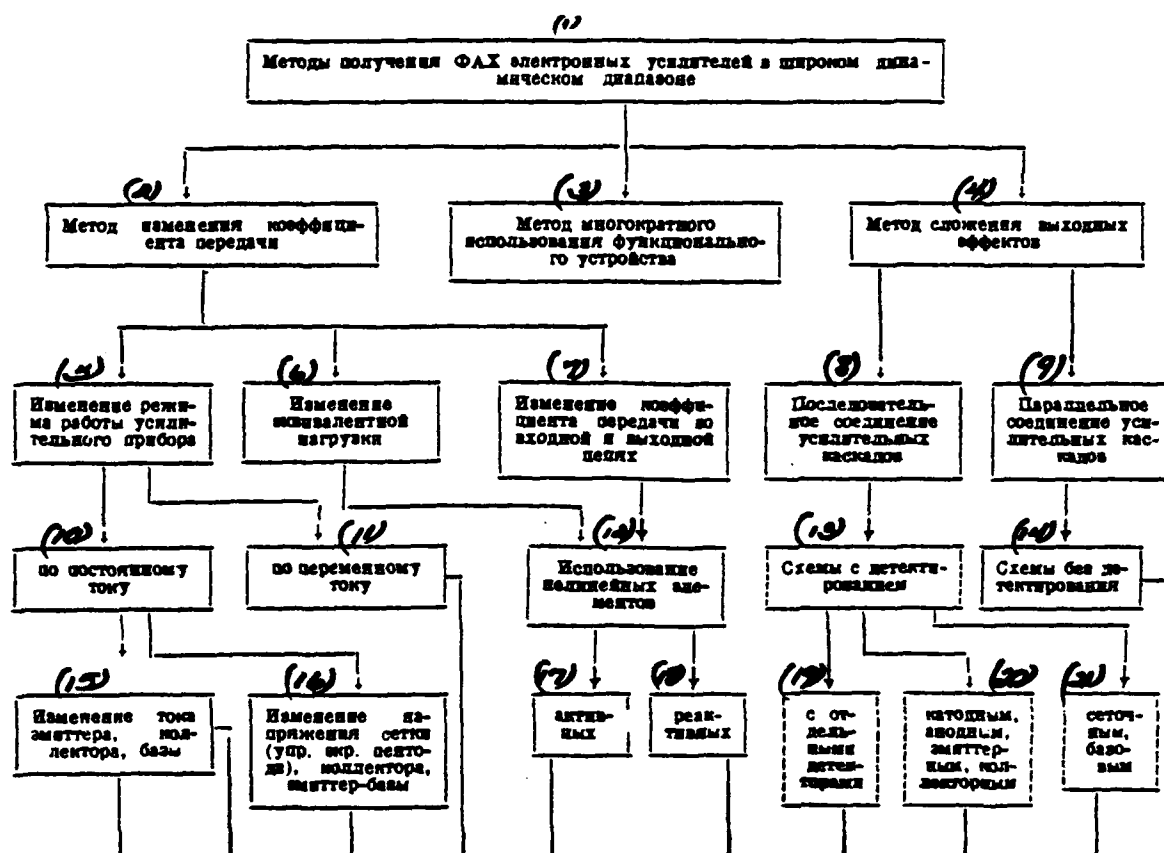
Expressions for the linear and logarithm-discontinuous characteristics are given in Table 1. Expressions for the remaining characteristics can be obtained by method indicated above.

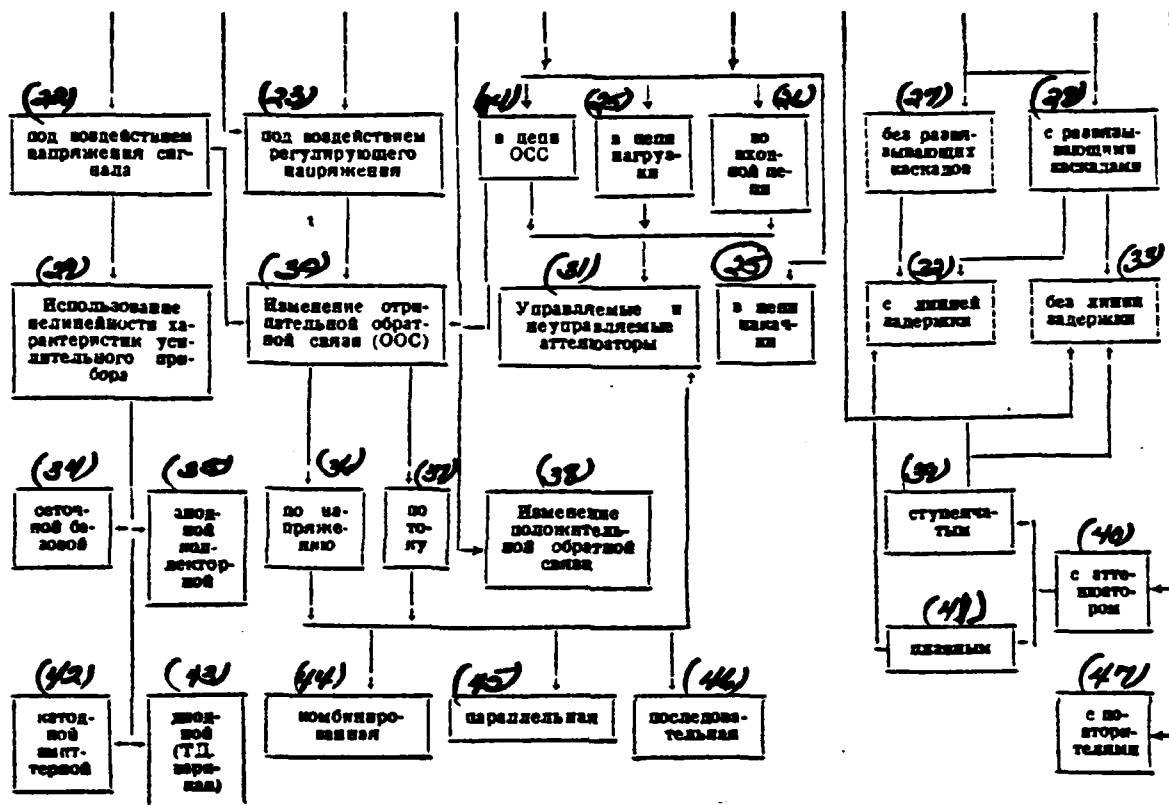
To realize FAKh in the broad dynamic band of a change of the input effect is possible only in the case of multistage amplifiers. There is a large number of circuit solutions of the realization of FAKh, which can be joined into the following three methods (Table 2): a change of the transmission factor; the addition of output effects; the repeated use of a functional unit (element/cell).

Latter/last method also can be realized on the base of the first two methods.

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Table 2.





Within the framework, encircled by solid line, the solutions for the aperiodic and selective pulse amplifiers are shown.

The solutions, suitable only for the selective amplifiers, they are shown within the dash framework.

Key: (1). Methods of obtaining FAKh of electronic amplifiers in the broad dynamic band. (2). Method of changing transmission factor. (3).

Method of repeated use of functional unit. (4). Method of adding output effects. (5). Change in mode/conditions of work of amplifier instrument. (6). Change in equivalent load. (7). Change in transmission factor in input and output circuits. (8). Series connection of amplifier stages. (9). Parallel connection of amplifier stages. (10). on direct current. (11). on alternating current. (12). Use by nonlinear of elements/cells. (13). Diagrams with detection. (14). Diagrams without detection. (15). Change in current of emitter, collector/receptacle, base. (16). Change in voltage/stress of grid (cont. screen pentode), collector/receptacle, emitter-base. (17). active. (18). reactive/jet. (19). with separate detectors. (20). cathode, anodic, emitter, collector. (21). grid, base. (22). under effect of voltage of signal. (23). under effect of controlling voltage/stress. (24). in circuit OSS. (25). in load circuit. (26). in input circuit. (27). without untying cascades/stages. (28). with untying cascades/stages. (29). Use of nonlinearity of characteristics of amplifier instrument. (30). Change in negative feedback (OOS). (31). Controlled and unguided attenuators. (32). with delay line. (33). without delay line. (34). by grid base. (35). by anodic collector. (36). on voltage/stress. (37). on current. (38). Change in positive feedback. (39). stepped. (40). with attenuator. (41). smooth. (42). by cathode emitter. (43). diode (TD varicap). (44). combined. (45). parallel. (46). consecutive. (47). with repeaters.

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§4. Method of changing the transmission factor.

Functional amplifier in the general case is nonlinear amplifier with the variable amplification factor. Therefore FAKh in the electronic amplifier it is possible to obtain by an automatic change of its amplification factor with the increase of input effect. The possible circuit solutions of the automatic gain control (ARU) are enumerated in Table 2.

It is necessary to note that FAKh in the broad dynamic band 80-100 dB in one nonlinear cascade/stage in practice is impossible to obtain. However, it is possible to obtain it in the amplifier, which consists of n nonlinear cascades/stages (Fig. 10). In this case the transmission factors in the nonlinear cascades/stages must be changed according to the completely specific laws.

Work [104] examines a special case of obtaining FAKh and it is shown that it is possible to obtain approximately/exemplarily LAKh of n - cascade amplifier in ShDD, if each of the cascades/stages has the amplitude characteristic, which consists of two linear sections, described by equations (1-56), in which coefficient $b_0=1$. General/common/total AKh of amplifier consists of the individual

sections, whose ends/leads lie/rest on the logarithmic curve. However, the isolated points of objective parameter considerably differ from logarithmic curve. For an increase in the accuracy of LAKh it is necessary to reduce the factors of amplification of cascades/stages and to increase a number of cascades/stages, which economically and is structurally/constructurally disadvantageous.

Consequently, for obtaining precise FAKh amplifier stages of multistage FU must be nonlinear with the completely specific amplitude characteristics, whose character is determined by the mode of operation of cascades/stages. Are possible the following modes of operation of the nonlinear cascades/stages: the strictly successive work of nonlinear cascades/stages in the functional modes/conditions; the simultaneous work of two or several cascades/stages (pairs, sets of three, etc.) in nonlinear, but nonfunctional modes/conditions (combined method of obtaining FAKh).

The mode of operation of cascades/stages is determined by character of FAKh and diagram of gain control.

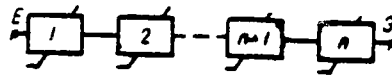


Fig. 10. Block diagram of n-cascade FU with the nonlinear cascades/stages.

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Analysis of the successive work of nonlinear cascades/stages.

Let us consider the work of nonlinear cascades/stages on the block diagram of n-cascade amplifier (Fig. 10) and let us determine the requirements, by which we must satisfy nonlinear cascades/stages for obtaining precise FAKh in ShDD. Since these requirements do not depend on the method of adjustment and form of the load of cascade/stage (aperiodic or selective load), analysis can be carried out, not taking into account the concrete/specific/actual circuit solution.

We accept the following designations:

$F(E)$ - FAKh of n-cascade amplifier, which must be realized;

$f_i(v_i)$ - amplitude characteristic of the i nonlinear cascade/stage;

E and v_i - effect respectively at the input of amplifier and i nonlinear cascade/stage;

E, v_i - output effects of amplifier and i cascade/stage.

The amplitude characteristic of amplifier, on one hand, according to expression (1-24), can be registered so:

$$g = F(E) = F(E_0) + \Delta\Phi(E); \quad (1-63)$$

on the other hand as the characteristic, formed/shaped with n by the series-connected nonlinear cascades/stages,

$$F(E) = f_n(v_n), \quad (1-64)$$

where

$$\left. \begin{aligned} v_n &= f_{n-1}(v_{n-1}); \\ v_{n-1} &= f_{n-2}(v_{n-2}); \\ &\dots \dots \dots \\ v_2 &= f_2(v_1); \\ v_1 &= f_1(v_1); \\ v_1 &= E, \end{aligned} \right\} \quad (1-65)$$

or in the convoluted form

$$F(E) = f_n \{ f_{n-1} [f_{n-2} \dots f_2 \{ f_1(E) \}] \}. \quad (1-66)$$

Amplitude characteristic of each cascade/stage according to expression (1-19)

$$f_i(v_i) = f_i(v_{i0}) + \Delta\phi_i(v_i). \quad (1-67)$$

Nonlinear cascades/stages can be identical and different.

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Successive operation of identical nonlinear cascades. In this case the equalities

$$K_{u1} = K_{u2} = \dots = K_{u, n-1} = K_{un} = K_u; \quad (1-68)$$

$$v_{u1} = v_{u2} = \dots = v_{u, n-1} = v_{un} = v_u; \quad (1-69)$$

$$d_1 = d_2 = \dots = d_{n-1} = d_n = d; \quad (1-70)$$

$$\Delta\phi_1(v_1) = \Delta\phi_2(v_2) = \dots = \Delta\phi_n(v_n) = \Delta\phi(E), \quad (1-71)$$

are implemented where $d_i = \frac{v_{ui}}{v_{ui}}$ - dynamic range of the functional section of the i cascade/stage.

Under some input influence, $E < E_n$, all cascades/stages work in the linear conditions and effect at the output of the amplifier

$$F(E) = f_n(v_n) = K_u^n v_1 = K_0 E, \quad (1-72)$$

where $K_0 = K_u^n$ - transmission factor of n -cascade amplifier with the work in the linear conditions.

Let us assume that when $E_1 = E_n$ the effect at the input of the n cascade/stage is equal to $v_n = v_{un}$ and latter/last n cascade/stage entered the functional mode/conditions. Since all previous cascades/stages work in the linear conditions,

and

$$v_{nn} = K_n^{n-1} E_n \quad (1-73)$$

$$F(E_n) = f_n(v_n) = K_0 E_n. \quad (1-74)$$

With change v_n from v_n to $v_n = v_n d$, which corresponds to change E from $E_1 = E_n$ to $E_2 = E_1 d$, latter/last cascade/stage works in the functional mode/conditions in the amplitude characteristic of amplifier it is formed/shaped with this cascade/stage, i.e.,

$$F(E) = f_n(v_n) = f_n(v_n) + \Delta \phi_n(v_n) = F(E_n) + \Delta \phi(E), \quad (1-75)$$

$$E_1 < E < E_2, v_n < v_n < v_n \quad v_n < v_n < v_n \quad E_1 < E < E_2$$

since they are implemented equality (1-71) and (1-74).

Thus, at the output of amplifier is reproduced the assigned functional dependence. Since all previous cascades/stages work in the linear conditions, they introduce no distortions into $F(E)$.

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When $v_n = v_n$, which corresponds E_1 , latter/last cascade/stage leaves from the functional mode/conditions. So that the general/common/total characteristic $F(E)$ would not be distorted, next-to-last $(n-1)$ stage at this moment must enter the functional mode/conditions, whose input effect

$$v_{n-1} = \frac{v_n}{K_n}$$

must be equally to v_n . This is implemented under the condition

$$K_n = d. \quad (1-76)$$

In this case

$$F(E_2) = f_n(v_n) = f_n(v_n) + \Delta\phi_n(v_n) = F(E_n) + \Delta\phi(E_2).$$

With change v_{n-1} from v_n to v_n , which corresponds to change E from E_1 to $E_2 = E_n d^n$, next-to-last cascade/stage works in the functional mode/conditions and

$$\begin{aligned} F(E) &= f_n(v_n) = f_n[f_{n-1}(v_{n-1})] = f_n[f_{n-1}(v_n) + \Delta\phi_{n-1}(v_{n-1})] = \\ &= f_n(v_n) + b_n(v_n - v_n) = f_n(v_n) + b_n\Delta\phi_{n-1}(v_{n-1}), \end{aligned} \quad (1-77)$$

where b_n — differential transmission factor of latter/last cascade/stage after its output from the functional operating mode.

Precise FAKh when $E_1 < E < E_2$ is described by the expression

$$\begin{aligned} F(E) &= f_n(v_n) + \Delta\phi(E). \\ E_1 &< E < E_2 \quad E_1 < E < E_2 \end{aligned} \quad (1-78)$$

Equating expressions (1-77) and (1-78), we obtain the following condition of the realization of precise FAKh with the work in functional mode/conditions of (n-1) cascade/stage:

$$\begin{aligned} \Delta\phi(E) &= b_n\Delta\phi_{n-1}(v_{n-1}). \\ E_1 &< E < E_2, \quad v_n < v_{n-1} < v_n. \end{aligned}$$

Whence

$$b_n = \frac{\Delta\phi(E)}{\Delta\phi_{n-1}(v_{n-1})}. \quad (1-79)$$

Since, according to expression (1-71) $\Delta\phi_{n-1}(v_{n-1}) = \Delta\phi(E)$,

$$b_n(v_n) = 1, \quad (1-80)$$

$v > v_n.$

i.e. the differential transmission factor of latter/last cascade/stage after output from the functional operating mode must be equal to one.

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Let us agree to call this mode of operation of cascade/stage quasi-linear, since the running transmission factor $K = \frac{v}{v_n} = \frac{f(v)}{f(v_n)}$ in the implementation of many types of FAKh is changed and approaches one when $v \gg v_n$.

Amplitude characteristic of cascade/stage with the work in the quasi-linear mode/conditions can be written thus:

$$f(v) = f(v_n) + \Delta\phi(v_n) + b(v - v_n) = f(v_n) + \Delta\phi(v_n) + v_n \left(\frac{v}{v_n} - 1 \right). \quad (1-81)$$

Thus, the amplitude characteristic of nonlinear cascade/stage with the successive work must consist of three sections: linear, functional and quasi-linear (Fig. 11).

When $E_2 = E_1 d^2$

$$F(E_2) = f_n(v_n) + 2\Delta\phi(v_n) = F(E_1) + 2\Delta\phi(v_n),$$

since are implemented equalities (1-71) and (1-74).

Analogously it is possible to show that in the case of identical nonlinear cascades/stages the effect at the output of amplifier is equal to:

when $E_3 = E_1 d^3$

$$F(E_3) = F(E_1) + 3\Delta\phi(v_n);$$

.

when $E_n = E_1 d^{n-1}$

$$F(E_n) = F(E_1) + (n-1)\Delta\phi(v_n);$$

when $E_{n+1} = E_1 d^n$, when the first cascade/stage leaves from the functional mode/conditions,

$$F(E_{n+1}) = F(E_1) + n\Delta\phi(v_n) = F(E_1) + \Delta\phi(E_{n+1}), \quad (1-82)$$

i.e. at the output of amplifier always is reproduced the assigned functional dependence (Fig. 12).

Thus, for obtaining precise FAKh of n-cascade amplifier with the strictly next work of identical cascades/stages each of the cascades/stages must satisfy the following requirements:

a) the amplitude characteristic of cascade/stage must consist of three sections: linear, functional and quasi-linear;

b) the dynamic range of FAKh of cascade/stage in the input effect must be equal to the initial transmission factor of cascade/stage (condition 1-76);

c) the differential transmission factor of nonlinear cascade/stage with the work in the quasi-linear mode/conditions must be equal to one.

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Greatest extent of the quasi-linear section of characteristic in latter/last cascade/stage. In the case of n of cascades/stages maximum effect at the output of latter/last cascade/stage toward the end of FAKh is determined by expression (1-82). In this case the maximum input effect, which still must be transmitted to the output with the differential coefficient of $b=1$,

$$\gamma_{nmax} = f_{n-1}(\gamma_{n-1m}) = f_{n-1}(\gamma_n) + (n-1) \Delta\phi(\gamma_n). \quad (1-83)$$

Dynamic range of FAKh of n -cascade amplifier in the case of the

identical cascades/stages

$$D = \frac{E_{n+1}}{E_n} = \frac{E_n d^n}{E_n} = d^n = K_n^n = K_0. \quad (1-84)$$

Analyzing expressions for the differential transmission factor $b(\nu)$, given in Table 1, it is possible to do the following conclusions:

1) with the successive work of the series-connected nonlinear cascades/stages it is possible to realize only FAKh with $b = \text{var}(\downarrow)$;

2) with the successive work of identical nonlinear cascades/stages it is possible to realize in the multistage amplifier only of LAKh with the foundation of logarithmic operation $N=e(a=1)$.

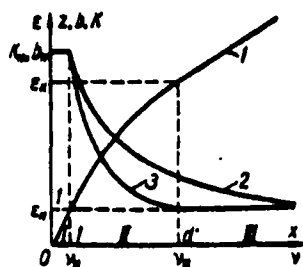


Fig. 11.

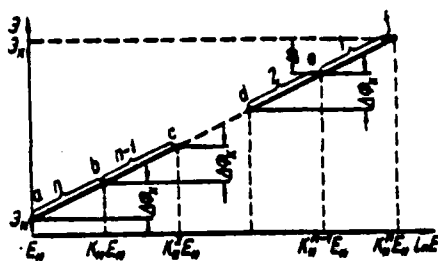


Fig. 12.

Fig. 11. Characteristics of nonlinear cascade/stage: 1 - $\epsilon=f(\nu)$, $z=f(x)$; 2 - $K=\psi(\nu)$; 3 - $b=0(\nu)$.

Fig. 12. Functional amplitude characteristic (LAKh) of multistage amplifier on semilogarithmic scale.

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If signal takes the logarithm of with the logarithm to the base N , different from e , then the differential transmission factor of cascade/stage at the end of the range of LAKh when $\nu = \nu_n = \nu_n = K_n$, as can be seen from Table 1, is equal to coefficient $b_n = a = \frac{1}{\ln N}$ and condition (1-80) virtually cannot be satisfied with the high accuracy, as a result of which general/common/total LAKh of n -cascade amplifier is distorted. For satisfaction of condition (1-80) it is necessary that the characteristic of cascade/stage at the point of

transition from the logarithmic section to the quasi-linear would have sharp fracture, what in practice cannot be carried out. In the objective parameter there will always be the smooth transition section, whose extent is greater, the more the foundation N differs from foundation e . But any divergence of the real amplitude characteristic of nonlinear cascade/stage distorts LAKh of amplifier.

With $a \neq 1$ according to (1-24) we have

$$F(E) = F(E_n) + a\Delta\Phi_n(E) = F(E_n) + \Delta\Phi_N(E). \quad (1-85)$$

If latter/last cascade/stage is supplied in the mode/conditions of logarithmic operation with foundation $N(a \neq 1)$, and all previous cascades/stages - into the mode/conditions of logarithmic operation with the Napierian base e , we will have:

$$\begin{aligned} f_n(v_n) &= f(v_n) + a\Delta\phi_{n,e}(v_n); \\ v_n &< v_n < v_n \\ f_i(v_i) &= f(v_n) + \Delta\phi_{i,e}(v_i). \end{aligned} \quad (1-86)$$

For the realization of dependence (1-85) according to expression (1-79) the differential coefficient of the transmission of latter/last cascade/stage with the work in the quasi-linear mode/conditions it must be equal to

$$b_n = \frac{a\Delta\Phi_n(E)}{\Delta\phi_{e,n-1}(v_{n-1})} = a, \quad (1-87)$$

since

$$\Delta\phi_{n-1,e}(v) = \Delta\Phi(E).$$

Then function $F(E_i)$ when E_i

$$F(E_i) = f_n(v_n) + (i-1) a \Delta \phi_n(v_n) = F(E_n) + (i-1) \Delta \phi_n(v_n), \quad (1-88)$$

$$E_i = E_n d^{i-1}$$

i.e., at the output of amplifier is reproduced the assigned dependence.

The standardized/normalized characteristics of nonlinear cascade/stage at the different values of coefficient of a (unbroken curves) are shown in Fig. 13.

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Thus, for obtaining precise LAKh of n-cascade amplifier with any foundation of logarithmic operation N it is necessary latter/last nonlinear cascade/stage to supply in the mode/conditions of logarithmic operation with the logarithm to the base N, and all remaining cascades/stages - into the mode/conditions of logarithmic operation with the Napierian base e.

Expressions for the amplitude characteristics and increment $\Delta \phi(E_n)$ in the nonlinear cascade/stage with the work in the quasi-linear mode/conditions are given in Table 3. Using these

expressions, and also formula (1-26), (1-27) and (1-85), we find the values of the dynamic range of n-cascade amplifier through the input effect and the contraction coefficient

$$D_{\text{max}} = \frac{F(E_n)}{F(E_n)} = a \ln d^n + 1 = na \ln d + 1; \quad (1-89)$$

$$C = \frac{d^n}{na \ln d + 1}. \quad (1-90)$$

The successive work of nonidentical nonlinear stages is caused by the fact that for some types of FU two conditions $b(v_n)=1$ and $K_n=d$ are not satisfied simultaneously. Thus, for instance, for the amplifiers with power characteristic $\epsilon = \alpha v^{\frac{1}{n}}$ condition $b(v)=1$ is satisfied when $d > K_n$.

For fulfilling the successive work of different nonlinear cascades/stages with different values K_{ni} or v_{ni} , but identical functional dependence $\Delta\phi(v_i)$ it is necessary that are fulfilled condition (1-80) and equality

$$f_{i-1}(v_{ni-1}) = v_{ni}, \quad (1-91)$$

or

$$v_{ni-1} K_{ni-1} = v_{ni} d_i. \quad (1-91a)$$

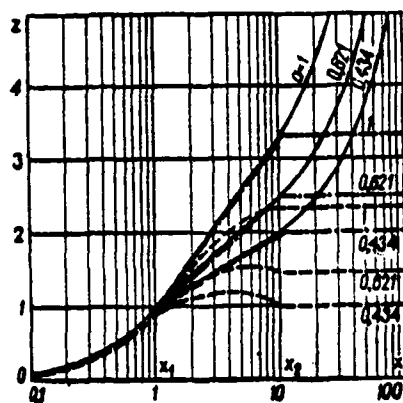


Fig. 13. Standardized/normalized amplitude characteristics of nonlinear cascade/stage on the semilogarithmic scale.

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On the basis of these conditions and with the fulfillment of equality (1-71), it is possible to carry out a successive work of two types of the different cascades/stages:

1. All nonlinear cascades/stages have identical initial transmission factor with the work in the linear conditions and different dynamic ranges of the functional section of the characteristic

$$K_{u1} = K_{u2} = \dots = K_{un} = K_u. \quad (1-92)$$

Then

$$y_{n+1} = \frac{y_{n1}}{K_{n+1}} = y_{n1} \frac{d_1}{K_u}$$

i.e. nonlinear cascades/stages must be distinguished by the level of initial input effect v_{ni} on which it begins with FAKh of cascade/stage.

2. Nonlinear cascades/stages have identical values v_n and different values d_i

$$v_{n1} = v_{n2} = \dots = v_{ni} = \dots = v_{nn} = v_n. \quad (1-93)$$

Then

$$K_{n1-i} = \frac{f_{i-1}(v_n)}{v_n} = \frac{v_{ni}}{v_n} = d_i, \quad (1-94)$$

i.e. the initial transmission factor of the previous cascade/stage with the work in the linear conditions must be equal to the dynamic range of FAKh of the subsequent cascade/stage.

Table 3.

Характеристика	Полная запись $\epsilon = f(\nu)$	Нормированная характеристика $\epsilon = f(x)$	$\Delta\phi(\epsilon_H)$	
$N = \epsilon$ ($a=1$)	$\epsilon = K_H \nu_H \times$ $\times \left(\ln d + \frac{\nu}{\nu_H d} \right)$	$\ln d + \frac{x}{d}$	$K_H \nu_H \ln d$	
ЛАХ	$N(a \neq 1)$	$\epsilon = K_H \nu_H \left[a \left(\ln d + \frac{\nu}{\nu_H d} - 1 \right) + 1 \right]$	$a \left(\ln d + \frac{x}{d} - 1 \right) + 1$	$K_H \nu_H a \ln d$
САХ $\beta = \frac{1}{n}$	$\epsilon = K_H \nu_H \left[(n-1) \times \right.$ $\times K_H^{\frac{1}{n-1}} + \frac{\nu}{\nu_H K_H} - n + 1 \left. \right]$	$(n-1) K_H^{\frac{1}{n-1}} +$ $+ \frac{x}{K_H} - n + 1$	$K_H \nu_H n (\sqrt[n]{d} - 1)$	

Key: (1). Characteristic. (2). Complete recording. (3).
Standardized/normalized characteristic.

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For n-cascade amplifier initial transmission factor and dynamic range is respectively equal to

$$K_0 = \prod_{i=1}^n K_{Hi} \quad (1-85)$$

$$D = \prod_{i=1}^n d_i. \quad (1-86)$$

The particular example to the realization of precise FAKh with the successive work of nonlinear cascades/stages can be realization of SAKh with $\beta < 1$. As can be seen from Table 1, equality $b(\nu) = 1$ is

implemented under standardized/normalized input influence x_n , equal to dynamic range of SAKh of cascade/stage, i.e.

$$x_n = d = K_n^{\frac{n}{n-1}}. \quad (1-97)$$

On the basis of expressions (1-19), (1-31), (1-81) and (1-97) it is easy to obtain expression for the quasi-linear section of the characteristic of nonlinear cascade/stage and increment $\Delta\phi(E_n)$, which are given in Table 3.

Let us determine the parameters of n-cascade FU with the power characteristic at satisfaction of conditions (1-92) and (1-93).

Condition (1-92) is satisfied. If the initial transmission factor of latter/last i cascade/stage and the input effect, under which it begins with its FAKh, is respectively equal to K_n and v_{n1} , then input effect v_{n1-i} (i-1) cascade/stage, with which it begins with its FAKh, according to (1-93) and (1-97),

$$v_{n1-i} = v_{n1} \frac{d_i}{K_n} = v_{n1} K_n^{\frac{i}{n-1}}. \quad (1-98)$$

For (i-m) cascade/stage

$$v_{n1-m} = v_{n1} K_n^{\frac{m}{n-1}}. \quad (1-99)$$

The dynamic range of FAKh of cascade/stage on input effect is determined by expression (1-97), and on the output effect

$$d_{\max} = z_n = nK_{\text{st}}^{\frac{1}{n-1}} - n + 1. \quad (1-100)$$

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Initial transmission factor and the dynamic range of FAX of i-cascade amplifier

$$K_0 = K_{\text{st}}^i; \quad (1-101)$$

$$D = d^i = K_{\text{st}}^{\frac{in}{n-1}}; \quad (1-102)$$

$$D_{\max} = n \sqrt[n]{D} - n + 1. \quad (1-103)$$

Condition (1-93) is satisfied. In this case the initial transmission factor of the i cascade/stage is equal to K_{st} .

Then according to expressions (1-94), (1-97, (1-100) for the i cascade/stage we have

$$\left. \begin{aligned} d_i &= K_{\text{st}}^{\frac{n}{n-1}}; \\ d_{\max i} &= nK_{\text{st}}^{\frac{1}{n-1}} - n + 1; \end{aligned} \right\} \quad (1-104)$$

for (i-1) cascade/stage

$$\left. \begin{aligned} K_{\text{st} i-1} &= d_i = K_{\text{st}}^{\frac{n}{n-1}}; \quad d_{i-1} = K_{\text{st} i-1}^{\frac{n}{n-1}} = K_{\text{st}}^{\left(\frac{n}{n-1}\right)^2}; \\ d_{\max i-1} &= nK_{\text{st} i-1}^{\frac{1}{n-1}} - n + 1; \end{aligned} \right\} \quad (1-105)$$

for (i-m) cascade/stage

$$K_{m1-m} = K_{m1}^{\left(\frac{n}{n-1}\right)^{m-1}}; d_{1-m} = K_{m1}^{\left(\frac{n}{n-1}\right)^m}; d_{\text{BMAX } 1-m} = K_{m1-m} - n + 1; (1-106)$$

for the 1st cascade/stage

$$K_{m1} = d_1 = K_{m1}^{\left(\frac{n}{n-1}\right)^{1-1}}; d_1 = K_{m1}^{\left(\frac{n}{n-1}\right)^1}; d_{\text{BMAX } 1} = nK_{m1} - n + 1. (1-107)$$

Qualitative indices of i-cascade amplifier

$$K_0 = K_{m1}^h; (1-108)$$

$$D = K_{m1}^{\frac{n}{n-1} h}; (1-109)$$

$$D_{\text{BMAX}} = nK_{m1}^{\frac{h}{n-1}}, (1-110)$$

where

$$h = \left[1 + \frac{n}{n-1} + \frac{n^2}{(n-1)^2} + \dots + \left(\frac{n}{n-1} \right)^{i-1} \right].$$

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Successive work of pairs, sets of three and so forth of nonlinear cascades/stages.

With the successive work of pairs or sets of three of cascades/stages each pair (sets of three) must have the amplitude characteristic, which satisfies the requirements of successive work in the functional mode/conditions. Consequently, AKh of pair (set of three) must also consist of three sections: linear, functional and

quasi-linear.

Let us agree on subsequently two (three) nonlinear of cascade/stage, which are component element FU, to in abbreviated form call pair (set of three), and all values, which relate to them, to accompany by indices "p" or "t".

With the work of pair (set of three) in the functional mode/conditions the cascades/stages of pair work in the nonlinear, but not functional mode/conditions.

It is analogous, as with the successive work of cascades/stages, pairs (set of three) can be identical and different. On the identical pairs (sets of three) can be carried out only the amplifier with LAKh with the foundation of logarithmic operation e , while on the different ones - any FU with $b = \text{var}(\downarrow)$.

Amplifier stages in the pairs (sets of three) can be connected in series (Fig. 14a, b) or in parallel (Fig. 14c, d).

Upon the parallel connection of cascades/stages in the pair it is possible to realize the quasi-linear section of the amplitude characteristic of the pair of considerable range, and consequently, to realize precise FAKh in the broad dynamic band.

The amplitude characteristic of consecutive pair can be registered

$$f_n(\nu_n) = f_2[f_1(\nu_1)] = f_2[f_1(\nu_n)], \quad (1-111)$$

where $f_1(\nu_1)$ and $f_2(\nu_2)$ — amplitude characteristics of the second and first cascades/stages.

Changing to the standardized values, can be registered

$$e_n = f(\nu_n) = K_{n,n} \nu_{n,n}^2. \quad (1-112)$$

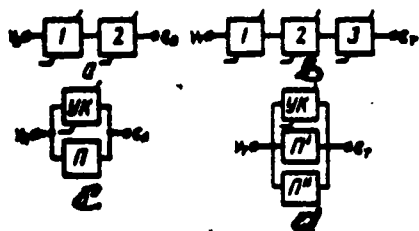


Fig. 14. Block diagrams of pairs and sets of three with the connection of the cascades/stages: a), b) — consecutive; c), d) — parallel.

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Running transmission factor of the pair

$$K_u = \frac{e_n}{v_u} = K_{u, n} \frac{z_n}{x_u}, \quad (1-113)$$

where $z_u = \frac{e_n}{e_{n, n}}$ and $x_u = \frac{v_n}{v_{n, n}}$.

Let us consider consecutive and parallel pair briefly.

Consecutive pair can consist:

1) of two nonlinear amplifier stages with the identical initial transmission factors

$$K_{u1} = K_{u2} = \sqrt{K_{u, n}}. \quad (1-114)$$

Then

$$v_{u1} = v_{u, n}; \quad v_{u2} = v_{u1} K_{u1} = v_{u, n} \sqrt{K_{u, n}}. \quad (1-115)$$

Running transmission factor of each cascade/stage of the pair

$$K_1 = K_2 = \sqrt{K_u} = \sqrt{K_{u, n} \frac{z_n}{x_u}}; \quad (1-116)$$

2) from the nonlinear attenuator from $K_u = 1$ and nonlinear amplifier stage.

Upon the inclusion/connection of the attenuator before amplifier stage the latter works in the lightened mode/conditions over the dynamic range of input signal.

In this case:

$$\left. \begin{aligned} K_{u1} &= 1; \\ K_{u2} &= K_{u. n}; \end{aligned} \right\} \quad (1-117)$$

$$v_{u1} = v_{u2} = v_{u. n}. \quad (1-118)$$

Running transmission factors of the cascades/stages

$$K_1 = \sqrt{\frac{z_n}{z_0}}; K_2 = K_{u. n} \sqrt{\frac{z_n}{z_0}}. \quad (1-119)$$

Upon the inclusion/connection of attenuator after amplifier stage are implemented the equalities

$$K_{u1} = K_{u. n}; K_{u2} = 1; v_{u1} = v_{u. n}; v_{u2} = v_{u1} = v_{u. n}. \quad (1-120)$$

Running transmission factors of the cascades/stages

$$K_1 = K_{u. n} \sqrt{\frac{z_n}{z_0}}; K_2 = \sqrt{\frac{z_n}{z_0}}. \quad (1-121)$$

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The amplitude characteristics of cascades/stages can be calculated by the formula

$$e = Kv.$$

Consecutive set of three can consist of different combinations of amplifier stages and attenuators. Consecutive set of three in practice is applied very rarely due to the complexity of tuning/adjusting.

Parallel pair (sets of three) just as consecutive pair can consist of different combinations of nonlinear amplifier stages and attenuators. However, pair [7], proposed by the author and which consists of nonlinear amplifier stage (UK) and linear repeater (P) (Fig. 14c) with the transmission factor, equal to one with a change of the input effect in ShDD, is great spread.

In general form AKh of parallel pair can be registered

$$f_3(v_n) = f_n(v_n) = f_1(v_n) + f_2(v_n) = f_1(v_n) + v_n, \quad (1-122)$$

where $f_1(v_n)$ — the amplitude characteristic of nonlinear amplifier stage (curve 3 in Fig. 15);

$f_2(v_n) = v_n$ — the amplitude characteristic of repeater (straight line 2 in Fig. 15).

Hence, the required characteristic of UK

$$f_1(v_n) = f_n(v_n) - v_n. \quad (1-123)$$

Characteristic $f_n(v)$ is described by expressions (1-24) and

(1-81).

If we into expression (1-123) substitute value $f_n(v_n)$ according to (1-81), we will have

$$f_1(v) = f(v_n) + \Delta\phi(v_n) - v_n = \text{const}, \quad (1-124)$$

$$v > v_n$$

i.e. with the work of pairs in the quasi-linear mode/conditions of the output effect UK must be constant. In this case quasi-linear section is formed/shaped with repeater with the differential transmission factor $b_1=1$.

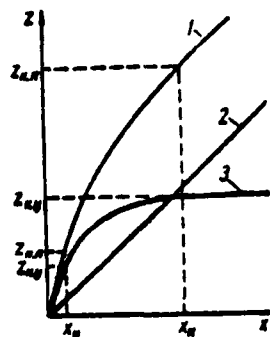


Fig. 15. The calibrated by FAKh parallel pairs on the graphic scale:
1 - pair; 2 - repeater; 3 - nonlinear amplifier stage.

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Table 4.

(1) Тип ФАХ		(2) Режим пары			
		(3) Функциональный режим $x_n < x < n$		(4) Квазилинейный режим $n < x < x_m$	
		(5) Полная запись $\epsilon = f(v)$	(6) Нормированная характеристика $\epsilon = f(x)$	(5) Полная запись $\epsilon = f(v)$	(6) Нормированная характеристика $\epsilon = f(x)$
ЛАХ, $\alpha = 1$	(7) Усилитель	$\epsilon = K_n v_n \left(\ln \frac{v}{v_n} + 1 \right) - v$	$\ln x + 1 - \frac{x}{K_n}$	$\epsilon = K_n v_n \ln d$	$\ln d$
	(8) Повторитель	$\epsilon = v$	$\frac{x}{K_n}$	$\epsilon = v$	$\frac{x}{K_n}$
САХ, $\beta = \frac{1}{n}$	(7) Усилитель	$\epsilon = K_n v_n \left(n \sqrt{\frac{v}{v_n}} - n + 1 \right) - v$	$n \sqrt{x} - n + 1 - \frac{x}{K_n}$	$\epsilon = K_n v_n \left(n \sqrt{\frac{d}{v_n}} - n + 1 - \frac{d}{K_n} \right)$	$n \sqrt{d} - n + 1 - \frac{d}{K_n}$
	(8) Повторитель	$\epsilon = v$	$\frac{x}{K_n}$	$\epsilon = v$	$\frac{x}{K_n}$

Key: (1). Type FAKh. (2). Mode/conditions of pair. (3). Functional mode/conditions. (4). Quasi-linear mode/conditions. (5). Complete recording. (6). Standardized/normalized characteristic. (7). Amplifier. (8). Repeater.

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As can be seen from the given analysis, it is expedient to apply the parallel pair (set of three), when are used amplifier instruments with a small dynamic range on the input effect, in particular, transistors. Connecting in parallel to nonlinear UK one, two repeaters, it is possible to obtain the quasi-linear section of the amplitude characteristic of the pair (set of three) of considerable range, and consequently, to carry out a successive work of a large number of nonlinear cascades/stages and to realize FAKh in ShDD.

At the given value of the initial transmission factor of pair $K_{n,n}$ the initial transmission factor of amplifier stage

$$K_{n,n} = K_{n,n} - K_1 = K_{n,n} - 1. \quad (1-125)$$

Expressions for amplitude characteristics on the voltage/stress of amplifier stage and repeater in the case of realization of LAKh with $a=1(N=e)$ and SAKh with $\beta < 1$ are given in Table 4. For the realization of LAKh with $N > e$ it is necessary latter/last pair to supply in the mode/conditions of guarantee $N > e$, and all previous pairs - into the mode/conditions of guarantee $N = e$.

Simultaneous work of n nonlinear cascades/stages.

In this case all n cascades/stages work in the nonlinear, but not functional mode/conditions. During this mode/conditions of cascades/stages it is possible to realize any FAKh, including with $b = \text{var}(\uparrow)$.

Let us consider the most real case, when all cascades/stages with the work in the linear conditions have identical initial transmission factor, i.e.

$$K_{n1} = K_{n2} = \dots = K_{nn} = K_n. \quad (1-126)$$

Running transmission factor of the amplifier

$$K = \frac{P(B)}{B} = K_n \frac{Z}{Y}, \quad (1-127)$$

where $K_n = K_n^*$.

Running transmission factor of cascade/stage with the work in the nonlinear mode/conditions

$$K_1 = K_1 = \dots = K_n = \sqrt[n]{K} = K_n \left(\frac{Z}{Y} \right)^{\frac{1}{n}}. \quad (1-128)$$

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Amplitude characteristic of the i cascade/stage

$$u_i = v_i K_i, \quad (1-129)$$

where v_i — the running voltage on the input of the i cascade/stage

$$v_i = EK_i^{i-1} = EK_n^{i-1} \left(\frac{Z}{Y} \right)^{\frac{i-1}{n}}. \quad (1-130)$$

Taking into account equalities (1-188) and (1-130) expression (1-189) can be registered in the form

$$e_i = EK_n^i \left(\frac{Z}{X} \right)^{\frac{i}{n}}. \quad (1-131)$$

In this case normalized AKh of the i nonlinear cascade/stage

$$z_i = \frac{e_i}{e_{ni}} = Z^{\frac{i}{n}} X^{\frac{n-i}{n}}, \quad (1-132)$$

where $e_{ni} = EK_n^i$; $X = \frac{E}{E_n}$; $Z = \frac{F(E)}{F(E_n)} = \frac{\partial}{\partial E}$ — normalized input effect and the output effect of amplifier.

§5. Method of adding the output effects.

Functional amplifier consists of n consecutively/serially (Fig. 16a) or in parallel (Fig. 16b) of connected amplifier stages (1, 2, ..., n), whose outputs are connected to the summator of effects. With the increase of the input effect FU amplifier stages, beginning from the latter, alternately leave from the linear conditions and enter into the saturation mode (limitation). The effects, which enter the summator from the value and do not depend on the level of the input effect of FU. Depending on type of UK and diagrams of their inclusion/connection the effect, removed from the output of summator and equal to the sum of the output effects of all amplifier stages, is found in one or the other functional dependence on the input effect of amplifier $\partial = F(E)$.

Amplifier stages can be linear (up to the saturation) (Fig. 16e) or nonlinear, identical or different. Naturally, if amplifier stage linear, precise FAKh cannot be obtained.

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For increasing the accuracy of FAKh in the diagram are connected the further corrective elements/cells (1, 2, ..., n in Fig. 16e), as which can be used the detectors (in the case of amplifying the radio signals), which untie cascades/stages, etc. If it is necessary to obtain very high accuracy of FAKh, such further elements/cells can be connected, also, with nonlinear UK.

Let us agree the set linear or nonlinear UK and the further corrective element/cell to call the nonlinear amplifier nucleus (I, II, ... N in Fig. 16e).

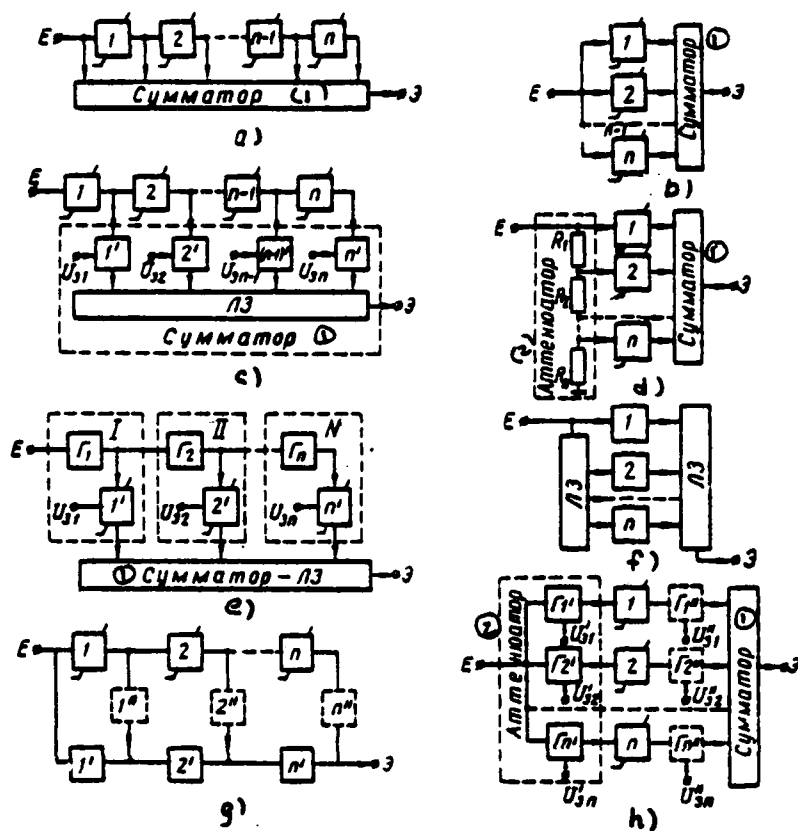


Fig. 16. Versions of block diagrams of FU, based on the methodology of consecutive addition voltage/stress.

Key: (1). Summator. (2). Attenuator.

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The circuit solution of summator can be most varied. In principle as the summator it is possible to use an active linear

resistor/resistance; delay line (LZ); separating cascades/stages with the linear amplitude characteristic in ShDD together with LZ (Fig. 16c) or without it; the repeater stages ($1'$, $2'$, ..., n' in Fig. 16g), included in parallel with UK, together with the separating cascades/stages ($1''$, $2''$, ..., n'') or without them. In the absence of separating cascades/stages of UK together with the repeaters are parallel pairs (S_4).

Upon the parallel connection of UK at the input of amplifier stepped or smooth attenuator (divider) can be connected. Stepped attenuator can consist of the series-connected active linear resistors/resistances (Fig. 16d), amplifier linear cascades/stages or amplifier-limiters; from the in parallel connected amplifier-limiters (Fig. 16h) with identical or different transmission factors with the work in the linear conditions.

As smooth attenuator LZ (Fig. 16f) can be used.

Let us agree FU upon the series connection of UK to call consecutive type FU, and upon the parallel connection of cascades/stages - parallel type FU. High dynamic quality is an advantage of FU of the consecutive type; deficiency/lack - change of phase-frequency response in the dynamic range. In parallel type FU, which have a comparatively small dynamic quality, it is possible to

obtain more stable phase-frequency response.

For determining the requirements, presented to UK and coordinating elements/cells, we analyze FU of both types.

Analysis of FU of the consecutive type.

Let us consider FU for different types of amplifier stages on the block diagram, depicted in Fig. 16,.

FU with the identical linear cascades/stages. Let us assume that all cascades/stages are identical, they work in the mode/conditions of linear amplification-limitation and have AKh , depicted in Fig. 17a; the transmission factor of cascade/stage with the work in the linear conditions is equal to K ; the coordinating elements/cells work in the linear conditions and have a transmission factor k .

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As a result of analysis it is necessary to determine the value of transmission factors k , for the realization of assigned $FAKh$.

Let us follow the work of amplifier with the gradual increase of the level of its input effect. On the small level E all

cascades/stages work in the linear conditions and effect at the output of the amplifier

$$\begin{aligned} \mathcal{O} = F(E) = E(K^n k_n + \\ + K^{n-1} k_{n-1} + \dots + K^2 k_2 + \\ + K k_1) = E \sum_{m=0}^{n-1} K^{n-m} k_{n-m}. \end{aligned} \quad (1-133)$$

Let us assume that under the input influence $E=E_1$ the latter/last n cascade/stage is saturated and from its output is taken the effect

$$e_{n1} = f_n(v_n) k_n = E_1 K^n k_n.$$

Effect at the output of the summator

$$\begin{aligned} \mathcal{O}_1 = E_1 K^n k_n + E_1 \sum_{m=1}^{n-1} K^{n-m} k_{n-m} = f_n(v_n) k_n + \\ + \sum_{m=1}^{n-1} f_{n-m}(v) k_{n-m}. \end{aligned} \quad (1-134)$$

With increase of E to E_1 from $E_2 = KE_1$ function $F(E)$ varies at output of summator linearly according to the law

$$\begin{aligned} F(E) = E_1 K^n k_n + E \sum_{m=1}^{n-1} K^{n-m} k_{n-m}. \end{aligned} \quad (1-135)$$

$$E_1 < E < E_2$$

The section of the general/common/total AKh of amplifier, described by expression (1-135), in Fig. 18a is designated by numbers 1-2. Index $(n-1)$ indicates that the section of 1-2 characteristics of amplifier is formed/shaped $(n-1)$ with cascade/stage.

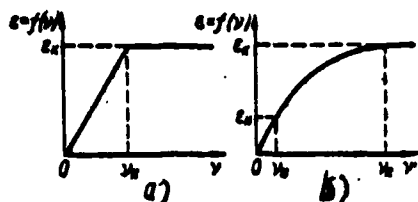


Fig. 17. Amplitude characteristics of UK: a) linear with the limitation; b) nonlinear.

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Under the input influence $E_n = KE_1$ is saturated next-to-last (n-1) cascade/stage and

$$\begin{aligned} \vartheta_n &= E_1 [K^n (k_n + k_{n-1}) + \sum_{m=2}^{n-1} K^{n-m+1} k_{n-m}] = \\ &= f(v_n) (k_n + k_{n-1}) + \sum_{m=2}^{n-1} f_{n-m}(v) k_{n-m}. \end{aligned} \quad (1-136)$$

since

$$f_n(v_n) = f_{n-1}(v_n) = f(v_n).$$

With $E_n = E_1 K^2$ is saturated (n-2) cascade/stage and

$$\begin{aligned} \vartheta_n &= E_1 (K^n \sum_{p=0}^2 k_{n-p} + \sum_{m=2}^{n-1} K^{n-m+2} k_{n-m}) = \\ &= f(v_n) \sum_{p=0}^2 k_{n-p} + \sum_{m=2}^{n-1} f_{n-m}(v) k_{n-m}. \end{aligned} \quad (1-137)$$

By analogy it is possible to register that when $E_{i+1} = E_1 K^i$, when is saturated (n-i) cascade/stage, effect at the output of the summator

$$\begin{aligned}
 \mathcal{G}_{i+1} &= E_1 (K^n \sum_{p=0}^i k_{n-p} + \sum_{m=i+1}^{n-1} K^{n-m+1} k_{n-m}) = \\
 &= f(v_N) \sum_{p=0}^i k_{n-p} + \sum_{m=i+1}^{n-1} f_{n-m}(v) k_{n-m}. \quad (1-138)
 \end{aligned}$$

The first cascade/stage is saturated when $E_n = E_1 K^{n-1}$. In this case

$$\mathcal{G}_n = \mathcal{G}_1 K^n \sum_{p=0}^{n-1} k_{n-p} = f(v_N) \sum_{p=0}^{n-1} k_{n-p}. \quad (1-139)$$

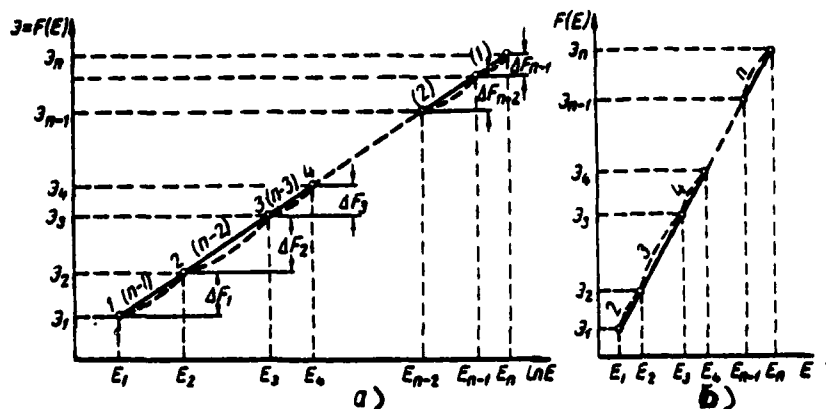


Fig. 18. Functional amplitude characteristics of amplifiers with the addition of the output effects: a) consecutive type amplifiers; b) parallel type.

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For the realization of the linear function $F(E)$ must be satisfied the condition

$$\frac{\partial z}{\partial E_1} = \frac{\partial z}{\partial E_2} = \dots = \frac{\partial z}{\partial E_n} = K \quad (1-140)$$

Condition (1-140) cannot be carried out in the case in question. However, it is possible to satisfy it, if to the corrective elements/cells we feed cutoff voltages.

For obtaining the logarithmic dependence $F(E)$ all points 1, 2, ...

3, ..., n at the semilogarithmic scale must lie/rest on the straight line. This will be in such a case, when is satisfied the condition of the equality of the slope/transconductance of the individual sections of characteristic (n-1, n-2, ..., 1) on the semilogarithmic scale, i.e.

$$\alpha_{n-1} = \alpha_{n-2} = \dots = \alpha_2 = \alpha_1 = \alpha. \quad (1-141)$$

According to expression (1-37)

$$\frac{\Delta \vartheta_1}{\ln d_{n-1}} = \frac{\Delta \vartheta_2}{\ln d_{n-2}} = \dots = \frac{\Delta \vartheta_{n-1}}{\ln d_1} = \alpha, \quad (1-142)$$

where $\Delta \vartheta_1 = \vartheta_2 - \vartheta_1$, $\Delta \vartheta_2 = \vartheta_3 - \vartheta_2$, ..., $\Delta \vartheta_{n-1} = \vartheta_n - \vartheta_{n-1}$; d_i — the working dynamic range of the i cascade/stage. In this case $d_i = \frac{E_{i+1}}{E_i} = K$.

Condition (1-42) can be rewritten

$$\Delta \vartheta_i = \alpha \ln d_{n-i+1}.$$

Solving system of equations (1-142) relatively k_i taking into account expressions (1-139), (1-138), (1-137) and so forth, we obtain

$$k_1 = \frac{\alpha \ln K}{E_1 K^{n-1} (K-1)} = \frac{\alpha \ln d}{E_1 K^{n-1} (K-1)} = \frac{\alpha \ln d}{E_n \left(1 - \frac{1}{K}\right)}; \quad (1-143)$$

$$k_2 = k_3 = \dots = k_{n-1} = k_n = k_1 \left(1 - \frac{1}{K}\right) = \frac{\alpha \ln d}{E_1 K^n} = \frac{\alpha \ln d}{E_n}. \quad (1-144)$$

Thus, for obtaining LAKh the transmission factors of all coordinating elements/cells, except the connected at the output first cascade/stage, must be equal to each other and are determined by expression (1-144). Since $E_1 K^n = \vartheta_1 - \vartheta_n$ — the output effect, with which it begins with LAKh, taking into account expression (1-144) can be registered:

$$\alpha = \frac{k E_1 K^n}{\ln d} = \frac{k \vartheta_n}{\ln K}. \quad (1-145)$$

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Thus, slope/transconductance LAKh depends on the transmission factors of cascades/stages and corrective elements/cells. With $K=e$ and $k=1$ the amplifier takes the logarithm of signal according to the law of natural logarithm.

For the realization of PAKh must be implemented, according to expression (1-46), the following condition:

$$\beta \ln M = \frac{\ln d_{\text{BMX } n-1}}{E_n - E_1} = \frac{\ln d_{\text{BMX } n-2}}{E_n - E_2} = \dots = \frac{\ln d_{\text{BMX } 1}}{E_n - E_{n-1}}, \quad (1-146)$$

where $d_{\text{BMX } n-1} = \frac{\mathcal{G}_1}{\mathcal{G}_n}; d_{\text{BMX } 1} = \frac{\mathcal{G}_n}{\mathcal{G}_{n-1}}.$

Taking into account that $(E_{i+1} - E_i) = E_i(d_i - 1)$ condition (1-146) it is possible to rewrite

$$\beta \ln M = \frac{\ln(\mathcal{G}_1 \mathcal{G}_1^{-1})}{E_1(d_{n-1} - 1)} = \frac{\ln(\mathcal{G}_2 \mathcal{G}_2^{-1})}{E_2(d_{n-2} - 1)} = \dots = \frac{\ln(\mathcal{G}_n \mathcal{G}_{n-1}^{-1})}{E_{n-1}(d_1 - 1)}. \quad (1-147)$$

Condition (1.147) in the case in question cannot be realized, but it is possible to realize it during the supplying of cutoff voltages on the corrective elements/cells and in parallel type amplifier.

For the realization of SAKh according to expression (1-53) must be satisfied the condition

$$\beta = \frac{\ln d_{\text{BMX } n-1}}{\ln d_{n-1}}. \quad (1-148)$$

For the present instance this condition can be registered thus:

$$\frac{g_{i+1}}{g_i} = \left(\frac{E_{i+1}}{E_i} \right)^{\beta} = K^{\beta}. \quad (1-149)$$

For finding n the transmission factors of the corrective elements/cells it is necessary to solve n equations with n unknowns. With the sufficiently high transmission factor, when the condition

$$K \gg 1, \quad (1-150)$$

is satisfied it is possible to consider the reaction only of the previous cascade/stage. In this case an error in the calculation is approximately/exemplarily equal to $1/K^2$.

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With the adopted assumption the transmission factors of the corrective elements/cells can be consecutively/serially calculated from the formula

$$k_{n-1} = \frac{K^{\beta} - 1}{1 - K^{\beta-1}} \sum_{m=0}^{n-1} k_{n-m}, \quad (1-151)$$

beginning with k_{n-1} at the given value β and k_n .

Let us find the value of the transmission factors of the corrective elements/cells, when on them cutoff voltages U , are given.

It is obvious that for the exception/elimination of effect of (i-1) cascade/stage on the form of the characteristic of amplifier with the formation of characteristic by the i cascade/stage the cutoff voltages must be equal to

$$U_{st} = v_{ni}. \quad (1-152)$$

FOOTNOTE 1. By v - it is necessary to understand the voltage of signal on the input of the i cascade/stage. ENDFOOTNOTE.

If cascades/stages all identical, voltage/stress U_{st} are equal to each other. On the latter/last corrective element/cell voltage/stress U_1 is not supplied. In this case it is possible to register:

$$\text{with } E < E_1 = \frac{v_n}{K^{n-1}}$$

$$P(E) = EK^n k_n = f_n(v_n) k_n; \quad (1-153)$$

$$\text{with } E = E_1$$

$$\partial_1 = E_1 K^n k_n = f_n(v_n) k_n; \quad (1-154)$$

$$\text{with } E_2 = KE_1$$

$$\begin{aligned} \partial_2 &= f_n(v_n) k_n + f_{n-1}(v_n) k_{n-1} - v_n = \\ &= E_1 [K^n k_n + (K-1) K^{n-1} k_{n-1}]; \end{aligned} \quad (1-155)$$

$$\text{with } E_3 = E_2 K = E_1 K^2$$

$$\begin{aligned} \partial_3 &= f_n(v_n) k_n + f_{n-1}(v_n) k_{n-1} - v_n + f_{n-2}(v_n) - v_n = \\ &= E_1 [K^n k_n + (K-1) K^{n-1} k_{n-1} + (K-1) K^{n-2} k_{n-2}]; \end{aligned} \quad (1-156)$$

with $E_i = E_i K^i$

$$\begin{aligned} \partial_i &= \sum_{m=0}^i f_{n-m}(v_n) k_{n-m} - 3v_n = \\ &= E_i K^{n-1} [K k_n + (K-1) \sum_{m=1}^i k_{n-m}]; \end{aligned} \quad (1-157)$$

with $E_{i+1} = E_i K^i$

$$\begin{aligned} \partial_{i+1} &= \sum_{m=0}^i f_{n-m}(v_n) k_{n-m} - i v_n = \\ &= E_i K^{n-1} [K k_n + (K-1) \sum_{m=1}^i k_{n-m}]. \end{aligned} \quad (1-158)$$

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Taking into account expression (1-140) and solving (1-154)-(1-158) relatively k_i , we obtain formulas for calculation k_i according to the given value of K and k_n in the implementation of the linear function $F(E)$

$$\left. \begin{aligned} k_{n-1} &= k_n K; \\ k_{n-2} &= k_n K + k_{n-1} (K-1); \\ k_{n-m} &= k_n K + (K-1) \sum_{i=1}^{m-1} k_{n-i}. \end{aligned} \right\} \quad (1-159)$$

Substituting consecutively/serially k_{i-1} in K_i , we obtain the simpler general/common/total expression for the transmission factor of $(n-m)$ corrective element/cell

$$k_{n-m} = k_n K^m. \quad (1-160)$$

Multistage linear amplifier can be realized, if to assign value $k_n \ll 1$.

Taking into account expression (1-142) and solving (1-154)-(1-158) relatively k_i , we obtain the following condition of realization of LAKh:

$$k_{n-1} = k_{n-2} = \dots = k_2 = k_1 = \frac{\sigma \ln K}{E_1 (K-1) K^{n-1}}. \quad (1-161)$$

Whence slope/transconductance of LAKh at the given values of K and k

$$\sigma = k E_1 (K-1) K^{n-1} (\ln K)^{-1}, \quad (1-162)$$

i.e. is obtained somewhat smaller in comparison with the slope/transconductance of LAKh in the absence of cutoff voltages (1-145).

Taking into account condition (1-147), (1-148), we obtain recursion formulas for calculation k_i in the implementation of the exponential and exponential functions

$$\left. \begin{aligned} k_{n-1} &= \frac{K k_n (0-1)}{K-1}; \\ k_{n-2} &= \frac{[K k_n + (K-1) k_{n-1}] (0-1)}{K-1}; \\ &\dots \dots \dots \\ k_{n-m} &= \frac{[K k_n + (K-1) \sum_{i=1}^{m-1} k_{n-i}] (0-1)}{K-1} \end{aligned} \right\} \quad (1-163)$$

where $\theta = M^{\theta E_1 (K-1)}$ — in the case of realization of PAKh; $\theta = K^\theta$ — in the case of realization of SAKh.

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If stages are identical, dynamic range of FAKh

$$D = K^{n-1}. \quad (1-164)$$

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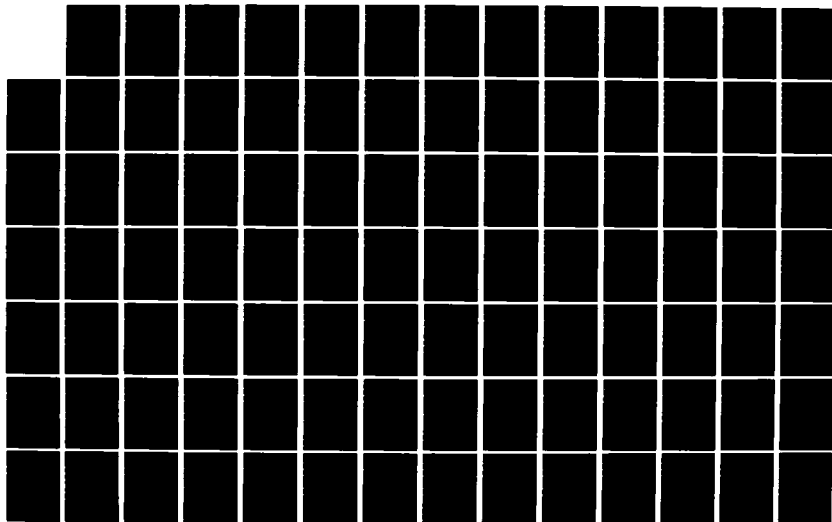
FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC
BAND(U) FOREIGN TECHNOLOGY DIV WRIGHT-PATTERSON AFB OH
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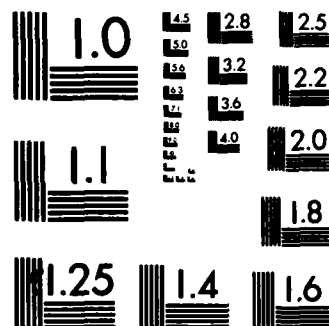
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It is necessary to note that in practice it is very difficult to perform amplifier from the identical cascades/stages in view of the scatter of the parameters of amplifier instruments. Therefore the case, when FU is made from different cascades/stages, is of practical interest.

FU with different linear cascades/stages. Let us assume that the cascades/stages have the amplitude characteristics, depicted in Fig. 17a, different transmission factors with the work in linear conditions K_i and different levels of input effect v_i , with which they enter into the mode/conditions of limitation.

1. To corrective elements/cells are given cutoff voltages, which according to expression (1-152) must be equal to

$$U_{en-1} = v_{nn}, U_{en-2} = v_{nn-1}, \dots, U_{e1-1} = v_{n1}.$$

With input effect $E_1 = \frac{v_{nn}}{\prod_{i=1}^n K_i}$, the saturation of latter/last n cascade/stage will occur. Effect at the output of the amplifier

$$g_1 = f_n(v_{nn}) k_n = E_1 K_n \prod_{i=1}^n K_i, \quad (1-165)$$

effect at input of $(n-1)$ cascade/stage

$$v_{nn-1} = \frac{v_{nn}}{K_{n-1}}. \quad (1-166)$$

Next-to-last $(n-1)$ cascade/stage will enter the mode/conditions of limitation when v_{nn-1} , i.e. with the increase of input effect d_{n-1}

times

$$d_{n-1} = \frac{v_{nn-1}}{v_{nn-1}} = \frac{v_{nn-1}}{v_{nn}} \cdot K_n = \gamma_{n-1} K_{n-1}, \quad (1-167)$$

where $\gamma_{n-1} = \frac{v_{nn-1}}{v_{nn}}$ — the coefficient of a difference in the input effects, under which two series-connected cascades/stages enter into the mode/conditions of limitation.

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Let us agree value d_{n-1} to call working dynamic range of (n-1) cascade/stage. Analogously for (n-m) cascade/stage

$$d_{n-m} = \frac{v_{nn-m}}{v_{nn-m+1}} \cdot K_{n-m} = \gamma_{n-m} K_{n-m}.$$

Under the input influence of amplifier $E_s = E_1 d_{n-1}$ is saturated (n-1) cascade/stage and

$$\vartheta_2 = \vartheta_1 + E_1 (d_{n-1}) k_{n-1} \prod_{i=1}^{n-1} K_i;$$

when $E_s = E_1 d_{n-1} = E_1 d_{n-1} d_{n-2}$ is saturated (n-2) cascade/stage and

$$\vartheta_3 = \vartheta_2 + E_1 d_{n-1} (d_{n-2} - 1) k_{n-2} \prod_{i=1}^{n-2} K_i;$$

with

$$E_s = E_1 d_{n-1} = E_1 d_{n-1} d_{n-2} d_{n-3}$$

$$\vartheta_4 = \vartheta_3 + E_1 d_{n-1} d_{n-2} (d_{n-3} - 1) k_{n-3} \prod_{i=1}^{n-3} K_i.$$

Taking into account condition (1-140), we obtain the following recursion formulas for calculation k_i in the implementation of the

linear function $F(E)$

$$\left. \begin{aligned} k_{n-1} &= k_n K_n; \\ k_{n-2} &= \frac{k_n K_n K_{n-1} + (d_{n-1} - 1) k_{n-1} K_{n-1}}{d_{n-1}}; \\ &\dots \dots \dots \end{aligned} \right\}$$

$$k_{n-m-1} = \frac{k_n \prod_{i=1}^m K_{n-i} + (d_{n-1} - 1) k_{n-1} \prod_{i=1}^m K_{n-i} + (d_{n-1} - 1) k_{n-2} \prod_{i=1}^m K_{n-i} + \dots + (d_{n-m} - 1) k_{n-m} \prod_{i=1}^{m-1} K_{n-i}}{\prod_{i=1}^m d_{n-i}}. \quad (1-168)$$

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Substituting consecutively/serially k_{n-1} in k_n , we obtain the simpler general/common/total expression for determining transmission factor $(n-m-1)$, which corrects element/cell at given values K_i and k_n .

$$k_{n-m-1} = k_n \prod_{i=1}^m K_{n-i}. \quad (1-169)$$

Taking into account condition (1-142) for the realization of LAKh we obtain

$$\left. \begin{aligned}
 k_{n-1} &= \frac{\sigma \ln d_{n-1}}{E_1 (d_{n-1} - 1) \prod_{i=1}^{n-1} K_i}; \\
 k_{n-2} &= \frac{\sigma \ln d_{n-2}}{E_1 d_{n-1} (d_{n-2} - 1) \prod_{i=1}^{n-2} K_i}; \\
 &\dots \dots \dots \\
 k_{n-m} &= \frac{\sigma \ln d_{n-m}}{E_1 (d_{n-m} - 1) \prod_{i=0}^{m-1} d_{n-i} \prod_{i=1}^{n-m} K_i}
 \end{aligned} \right\} (1-170)$$

Taking into account conditions (1-147) and (1-148) for the realization of PAKh and SAKh we obtain

$$\left. \begin{aligned}
 k_{n-1} &= \frac{\vartheta_1 (\theta_{n-1} - 1)}{E_1 (d_{n-1} - 1) \prod_{i=1}^{n-1} K_i}; \\
 k_{n-2} &= \frac{\vartheta_2 (\theta_{n-2} - 1)}{E_1 d_{n-1} (d_{n-2} - 1) \prod_{i=1}^{n-2} K_i}; \\
 k_{n-3} &= \frac{\vartheta_3 (\theta_{n-3} - 1)}{E_1 d_{n-1} d_{n-2} (d_{n-3} - 1) \prod_{i=1}^{n-3} K_i}; \\
 &\dots \dots \dots \\
 k_{n-m} &= \frac{\vartheta_m (\theta_{n-m} - 1)}{E_1 (d_{n-m} - 1) \prod_{i=1}^{m-1} d_{n-i} \prod_{i=1}^{n-m} K_i}
 \end{aligned} \right\} (1-171)$$

where $\theta_{n-m} = M^{E_1(d_{n-m}-1)}$ — in the case of realization of PAKh; $\theta_{n-m} = d_{n-m}^d$ — in the case of realization of SAKh.

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2. Corrective elements/cells work without cutoff voltages. In this case the slope/transconductance of formed/shaped FAKh is raised.

Let us register expressions for the output effect of amplifier $\partial_1, \partial_2, \dots, \partial_n$ on the levels of the input effect E_1, E_2 and so forth, the corresponding to the moments/torques of saturation $n_1(n-1)$ and so forth cascades/stages

$$\begin{aligned}\partial_1 &= E_1 (k_n \prod_{i=1}^n K_i + k_{n-1} \prod_{i=1}^{n-1} K_i + \dots + k_1 K_1); \\ \partial_2 &= E_1 [k_n \prod_{i=1}^n K_i + d_{n-1} (k_{n-1} \prod_{i=1}^{n-1} K_i + k_{n-2} \prod_{i=1}^{n-2} K_i + \dots + k_1 K_1)].\end{aligned}$$

The first cascade/stage is saturated under input influence

$$E_n = E_1 \prod_{i=1}^{n-1} d_{n-i}. \quad \text{In this case}$$

$$\begin{aligned}\partial_n &= k_n \prod_{i=1}^n K_i + d_{n-1} k_{n-1} \prod_{i=1}^{n-1} K_i + d_{n-1} d_{n-2} k_{n-2} \prod_{i=1}^{n-2} K_i + \\ &+ \dots + k_1 K_1 \prod_{i=1}^{n-1} d_{n-i}.\end{aligned}$$

Taking into account condition (1-142), we obtain the following recursion formulas for calculation k_i in the implementation of the logarithmic dependence $F(E)$:

$$\left. \begin{aligned}k_1 &= \frac{s \ln d_1}{E_1 E_1 (d_1 - 1) \prod_{i=1}^{n-1} d_{n-i}}; \quad k_2 = \frac{k_1 K_1}{\prod_{i=1}^1 K_i} \left[\frac{d_2 (d_1 - 1) \ln d_2}{(d_2 - 1) \ln d_1} \right]; \\ &\dots \dots \dots \\ k_m &= \frac{k_{m-1} \prod_{i=1}^{m-1} K_i + k_{m-2} \prod_{i=1}^{m-2} K_i + \dots + k_1 K_1}{\prod_{i=1}^m K_i} \left[\frac{d_m (d_{m-1} - 1) \ln d_m}{(d_m - 1) \ln d_{m-1}} \right].\end{aligned} \right\} \quad (1-172)$$

If into condition (1-148) to substitute complete

expanded/scanned values $d_{\text{max } n-1}$ will be obtained by n of equations with n by unknowns. As in the case of uniform cascades/stages, problem it is simplified, taking into account the output effect only of previous $(i-1)$ cascade/stage upon the saturation of the i cascade/stage. Are obtained the following recursion formulas for calculation k_i in the implementation of SAKh with $\beta < 1$

$$\left. \begin{aligned} k_{n-1} &= \frac{k_n K_n}{d_{n-1}}; \quad k_{n-2} = \frac{k_n K_n K_{n-1} + d_{n-1} k_{n-1} K_{n-1}}{d_{n-1} d_{n-2}}; \\ &\dots \dots \dots \\ k_n &\prod_{i=0}^{m-1} K_{n-i} + d_{n-1} k_{n-1} \prod_{i=1}^{m-1} K_{n-i} + \dots + \\ &\quad + K_{n-m+1} \prod_{i=1}^{m-1} d_{n-i} \\ k_{n-m} &= \frac{\quad}{\prod_{i=1}^m d_{n-i}}. \end{aligned} \right\} (1-173)$$

Overall dynamic range of FAKh on the input effect in the case of n different cascades/stages

$$D = \prod_{i=1}^{n-1} d_{n-i}. \quad (1-174)$$

In the implementation of FAKh in the amplifier, which consists of the linear cascades/stages, the objective parameter coincides with precise only by the isolated points, which correspond to the moments/torques of saturating the cascades/stages (point 1, 2, 3, ... in Fig. 18a). The intermediate points of objective parameter considerably differ from precise FAKh. This divergence depending on the value of the transmission factor K and working dynamic range d of cascade/stage.

Let us consider the section of 1-2 characteristics and it is defined, as accuracy FAKh depends on K and d . Let us assume that all cascades/stages are identical, without the corrective elements/cells and the summator linearly summarizes output effects ϵ from the transmission factors $k=1$.

Initial section 1-2 of FAKh on the graphic scale is depicted in Fig. 19. The maximum divergence of real FAKh $F_p(E)$ from precise $F_r(E)$ is observed between points 1-2, while the maximum divergence of differential transmission factor b (slope/transconductance) and the converted slope/transconductance σ - at points 1 and 2. For precise and real FAKh according to expression (1-24) and Fig. 19

$$\begin{aligned} \mathcal{F}_r &= F_r(E) = F(E_1) + \Delta\Phi(E); & (1-175) \\ \mathcal{F}_p &= F_p(E) = F(E_1) + (E - E_1) \operatorname{tg} \alpha, & (1-175a) \end{aligned}$$

where $F(E_1) = E_1 \operatorname{tg} \alpha_n = E_1 K_n$; $K_n = \operatorname{tg} \alpha_n$ - the initial transmission factor of amplifier.

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The slope/transconductance of precise FAKh at points 1 and 2 is respectively equal to

$$b_{\tau 1} = \frac{\partial F_{\tau}(E)}{\partial E_{E-E_1}} = \frac{\partial |\Delta \Phi(E)|}{\partial E_{E-E_1}} = \frac{\partial |\Delta \Phi(E_H + \Delta E)|}{\partial E_{\Delta E=0}}; \quad (1-176)$$

$$b_{\tau 2} = \frac{\partial |\Delta \Phi(E_1 - \Delta E)|}{\partial E_{\Delta E=0}} = \frac{\partial |\Delta \Phi(E_{1d} - \Delta E)|}{\partial E_{\Delta E=0}}. \quad (1-176a)$$

Slope/transconductance of real FAKh at points 1 and 2

$$b_p = \frac{\partial F_p(E_1 + \Delta E)}{\partial E_{\Delta E=0}} = \frac{\partial F_p(E_1 - \Delta E)}{\partial E_{\Delta E=0}} = \operatorname{tg} \alpha. \quad (1-177)$$

The maximum divergence of real FAKh from precise $\Delta \theta_m = \theta_{\tau} - \theta_p$ and input effect E_m with which is observed this divergence, it is possible to find from the condition

$$b_{\tau}(E_m) = b_p(E_m), \text{ or } \frac{\partial F_{\tau}(E_m)}{\partial E} = \frac{\partial F_p(E_m)}{\partial E} = \operatorname{tg} \alpha. \quad (1-178)$$

As a special case, let us find the divergences of the objective parameter of the logarithmic amplifier, made on the identical cascades/stages with the transmission factors K . Using expressions (1-136)-(1-138) and (1-175) and Fig. 19a, for precise and real LAKh it is possible to register

$$F_{\tau}(E) = F(E_1) + \frac{K^n E_1}{1 + u_d} \ln \frac{E}{E_1}; \quad (1-179)$$

$$F_p(E) = K^n E_1 + E \sum_{i=1}^{n-1} K^{n-i}, \quad (1-180)$$

where

$$F(E_1) = E_1 \sum_{i=0}^{n-1} K^{n-i}.$$

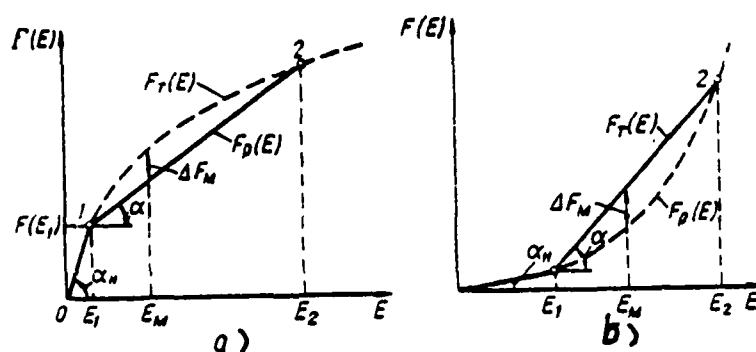


Fig. 19. Initial section of FAKh of multistage amplifier on the graphic scale: a) $b = \text{var} (\downarrow)$; b) $b = \text{var} (\uparrow)$; — real; ----- precise.

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On the basis of condition (1-178) we find that the maximum divergence

$$\Delta \vartheta_m = F(E_1) - \frac{K E_1}{\ln K} (\ln(A \ln d) + 1)$$

is observed under the input influence

$$E_m = \frac{K E_1}{A \ln d},$$

where

$$A = 1 + \frac{1}{K} + \frac{1}{K^2} + \dots + \frac{1}{K^{n-1}}.$$

Maximum relative deflection

$$\Delta \vartheta_m = \frac{\Delta \vartheta_m}{F_r(E_m)} = \frac{A - [\ln(A \ln d) + 1] (\ln K)^{-1}}{A + [1 - (\ln K)^{-1}] \ln(A \ln d)}. \quad (1-181)$$

According to expressions (1-176), (1-179) and (1-180) we find the values of the slope/transconductance of precise and real FAKh at points 1 and 2:

$$b_r = \frac{K^n E_1}{E \ln d}; \quad (1-182)$$

$$b_p = AK^{n-1}. \quad (1-183)$$

Then relative error on the slope/transconductance

$$\delta b = \frac{b_r - b_p}{b_r} = 1 - \frac{EA \ln d}{E_1 K}. \quad (1-184)$$

Slope/transconductance of LAKh in the converted coordinates is determined by expression (1-145). Let us find the expression for the slope/transconductance of objective parameter (1-180), examined/considered also in the semilogarithmic scale. For this let us introduce new variable $\alpha = \ln E$. Then expression (1-180) in the new coordinate system can be registered thus:

$$\partial_p = F_p(E) = K^n e^{\alpha} + e^{\alpha} \sum_{i=0}^{n-1} K^{n-i}.$$

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Slope/transconductance of the objective parameter

$$a_p = \frac{\partial \partial_p}{\partial \alpha} = e^{\alpha} \sum_{i=0}^{n-1} K^{n-i}.$$

Passing to variable E, we obtain

$$\sigma_p = \sum_{i=0}^{n-1} K^{n-i} = AK^{n-1}E. \quad (1-185)$$

Relative error on the slope/transconductance with $k=1$

$$\delta\sigma = \frac{\sigma_T - \sigma_p}{\sigma_T} = 1 - \frac{EA \ln d}{E_1 K}.$$

The obtained expression coincides with formula (1-184), i.e., relative error on the slope/transconductance of LAKh in the usual and converted coordinates coincides.

For point of inflection 1 with $E=E_1$ we have

$$\delta\sigma_1 = \frac{A \ln d}{K}, \quad (1-186)$$

for point of inflection 2 with $E=E_1=E_1 d$

$$\delta\sigma_2 = 1 - \frac{dA \ln d}{K}. \quad (1-186a)$$

Curves $\delta\sigma_1$, $\delta\sigma_2$ and $\delta\sigma_m = f(K)$, calculated by formulas (1-181) and (1-186) for case of $K=d$, are given in Fig. 20. During the calculation in coefficient of A were considered five members. From the curves it is evident that the slope/transconductance of objective parameter at point 1 is less, and at point 2 it is more than the slope/transconductance of precise LAKh at any value of transmission factor K . An error in the objective parameter increases with an increase in the transmission factor K and dynamic range d . Errors vanish with $d \rightarrow 1$ and $K \rightarrow 1$. However, with $K \rightarrow 1$ amplifier loses amplifier

properties.

Amplitude characteristic of real UK consists of the linear and nonlinear sections (Fig. 17b). With the transistorization nonlinear section can occupy significant part the characteristic of cascade/stage.

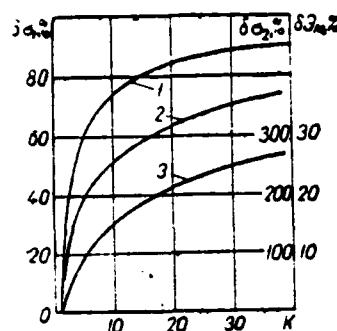


Fig. 20. The curves of the dependence of an error in real LAKh on the value of the transmission factor of cascade/stage K: 1 - $\delta\sigma_1$; 2 - $\delta\sigma_2$; 3 - $\delta\sigma_3$.

4

Page 65. If the character of the nonlinear characteristic of cascade/stage coincides with the character of formulated FAKh, then accuracy of FAKh can be considerably increased.

For obtaining FAKh of high accuracy nonlinear cascades/stages must have completely specific amplitude characteristics. Let us consider FU with the nonlinear cascades/stages.

Functional amplifiers with the identical nonlinear cascades/stages.

For obtaining precise FAKh of n-cascade amplifier it is necessary that combined AKh of each amplifier nonlinear cascade/stage and linear summator would satisfy the requirements of the successive

work of parallel pairs.

However, in this case there is a difference, consist in the fact that instead of the repeater in parallel to nonlinear cascade/stage is connected the summator, from which the summed up output effects of the previous cascades/stages do not enter the inputs of subsequent amplifier stages. This unidirectionality is ensured by the corrective elements/cells (detectors or the untying cascades/stages).

On one hand, the total characteristic of nonlinear cascade/stage and summator, which let us designate $f_2(v_i)$, it must be described by expressions (1-20), (1-24) and (1-81) and to satisfy conditions (1-76), (1-80), on the other hand, real total characteristic can be registered

$$f_{sp_i}(v_i) = k_i f_i(v_i) + \sum_{m=1}^{i-1} k_{i-m} f_{i-m}(v_{i-m}). \quad (1-187)$$

For obtaining precise FAKh must be implemented the equality

$$f_2(v_i) = f_{sp}(v_i). \quad (1-188)$$

Then the amplitude characteristic of the i nonlinear cascade/stage must be described by the expression

$$f_i(v_i) = \frac{1}{k_i} \left[f_2(v_i) - \sum_{m=1}^{i-1} k_{i-m} f_{i-m}(v_{i-m}) \right]. \quad (1-189)$$

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Taking into account that $f_{i-1}(v_{i-1}) = v_i$ and the initial transmission factors of all cascades/stages are equal, we can register

$$f_i(v_i) = \frac{1}{k_i} \left[f_i(v_i) - v_i \sum_{m=1}^{i-1} K^{1-m} k_{i-m} \right] \quad (1-189a)$$

For simplification in the construction/design of amplifier we accept $k_i = k_{i-1} = \dots = k_1 = 1$. Then expression (1-189) is simplified

$$f_i(v_i) = f_i(v_i) - v_i \sum_{m=1}^{i-1} K^{1-m}. \quad (1-190)$$

With the work in the linear conditions of i cascade/stage $v_i < v_{ni}$ and expression (1-190) can be rewritten

$$f_i(v_i) = v_i \left(K_i - \sum_{m=1}^{i-1} K^{1-m} \right) v_i K, \quad (1-191)$$

where K_i — initial transmission factor of function $f_i(v)$.

From expression (1-191) we obtain equation for determining the initial transmission factor of the i nonlinear cascade/stage

$$K = K_i - \sum_{m=1}^{i-1} K^{1-m}, \quad (1-192)$$

where m - number of the previous cascade/stage, whose output effect is considered during calculation K . To count off m necessarily from the i cascade/stage to the input of amplifier. With $m=1$ we obtain expression (1.125). With $m=2$ we obtain the equation of the second power

$$K^2 - K(K_2 - 1) + 1 = 0.$$

Whence

$$K = \frac{K_2 - 1}{2} + \sqrt{\frac{(K_2 - 1)^2}{4}} - 1. \quad (1-193)$$

An error in the calculation according to formula (1-192) can be approximately considered according to formula $\delta K = \frac{1}{K_2^m}$. Therefore with $K \geq 56$ K do not exceed 4%.

With $m=1$ the amplitude characteristics of nonlinear cascade/stage are described by the expressions, given in Table 4. At the arbitrary value of m of expression for AKh the cascades/stages will be analogous.

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The results, obtained in the previous paragraph during the analysis of the successive work of nonlinear cascades/stages, are suitable and in this case.

With the equality of the transmission factors of corrective elements/cells k , expression (1-186) can be registered

$$f_2(v_n) + \Delta\phi(v) = k[f_{2p}(v_n) + \Delta\phi_p(v)].$$

According to expression (1-37) $\Delta\phi(v) = a$, we can register

$$a = k\alpha_p. \quad (1-194)$$

Thus, changing the transmission factor k of the corrective elements/cells, it is possible to change the slope/transconductance of total FAKh. However, with a change of value k in the equal measure output effects $F(E_n)$ are changed and $F(E_n)$ dynamic range on the output effect of amplifier remains constant.

At constant value of $k=1$ this can be realized by inclusion/connection to the output FU of linear device/equipment with the transmission factor, equal to

$$k = \frac{a}{\alpha_p}. \quad (1-195)$$

In many instances it is necessary simultaneously with the decrease of slope/transconductance of FAKh to decrease range D_{max} . This can be realized as follows: latter/last cascade/stage together with the summator is placed in the mode/conditions of the guarantee of FAKh of the assigned slope/transconductance. Total characteristic is described by expressions (1-20), (1-24) and (1-81). The transmission factor of the latter/last corrective element/cell is taken by the equal to one, i.e., $k_n = 1$.

All remaining cascades/stages are placed in the most real mode/conditions with slope/transconductance σ_p , which more easily in all is ensured, and transmission factors k_i of the corrective elements/cells, connected at the outputs of these cascades/stages, are calculated from formula (1-195).

Then all results, obtained earlier, are valid for the cascades/stages, which precede the latter, for which initial transmission factor must be determined from the equation

$$K_n = k \sum_{m=1}^{n-1} K^{1-m} = K_1. \quad (1-196)$$

where K - initial transmission factors of cascades/stages; k - is determined according to formula (1-195).

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With $m=1$

$$K_n = K_1 - k = K_1 - \frac{a}{\sigma_p} \quad (1-197)$$

and AKh of the last cascade/stage must be described by the expression

$$f_n(v_n) = f_1(v) - \frac{a}{\sigma_p} v_n. \quad (1-198)$$

As an example, it is possible to consider obtaining LAKh with the foundation of logarithmic operation $N(a \neq 1)$. In this case

latter/last cascade/stage is placed in the mode/conditions of provision of LAKh with foundation N and is taken $k_n = 1$. All previous cascades/stages are placed in the mode/conditions of the guarantee of LAKh with foundation e(a=1) and the transmission factors of detectors they are taken by the following: $k_1 = a = \frac{1}{\ln N}$.

Then for the latter/last cascade/stage according to expressions (1-197) and (1-198)

$$K_n = K_1 - a;$$

$$z_{II} = a \ln x + 1 - \frac{ax}{K_1}; \quad z_{III} = a \ln K_n + 1 - \frac{aK_n}{K_1 - a}.$$

High accuracy of FAKh can be realized in the method of adding the output effects, if we on the outputs of cascades/stages supply the corrective elements/cells, the transmission factors of which are changed according to the law

$$k_i = \begin{cases} 0 & \text{при } v_{in} < f_i(v_{in}) = v_{in+i}; \\ K & \text{при } v_{in} > f_i(v_{in}) = v_{in+i}, \end{cases} \quad (1-199)$$

Key: (1). with.

where v_{in} — input effect of the i corrective element/cell.

In this case general/common/total FAKh of amplifier F(E) with the strictly successive work of nonlinear cascades/stages is formed/shaped without the effect of the output effects of the previous cascades/stages. Then AKh of cascades/stages when $v_1 < v_{in}$

they must be described by expressions (1-20), (1-24), and when $v_i > v_{th}$

$$f_i(v_i) = f_i(v_{th}) + \Delta\phi_i(v_{th}) = \text{const.} \quad (1-200)$$

Furthermore, must be satisfied condition (1-76). If in the objective parameter it is most easy to realize slope/transconductance ϕ , then value k is calculated from formula (1-195).

Law (1-199) can be realized during the supplying to the corrective elements of the cutoff voltages, whose value is determined by expression (1-152).

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However, to obtain the objective parameter of transmission with the sharp fracture is very difficult and it will have some section of smooth transition from $k_p = 0$ and $k_p = k$, which will introduce further error in FAKh. This section can be obtained quite small with small input voltage (for example, emitter detector) [8].

The supply of cutoff voltages on the corrective elements/cells considerably facilitates the adjustment of amplifier from the point of view of obtaining the assigned slope/transconductance of FAKh. However, in this case are their negative sides.

In the implementation of dependence (1-200) when $v < v_m$ must be observed sharp fracture in AKh the cascade/stage, which to carry out is virtually very difficult. The presence of transition section causes the divergence of the slope/transconductance of real FAKh from the slope/transconductance of precise FAKh at points E_n which correspond v_m . The mode/conditions examined it is expedient to apply in transistor FU, since the transistors more sharply pass into the saturation mode in comparison with vacuum lamps.

It is necessary to note that in the transistor amplifier it is very difficult to fulfill identical nonlinear cascades/stages.

FU with the different nonlinear cascades/stages.

Let us assume that FU consists of the nonlinear cascades/stages, which have different ones $f_1(v_1)$, K_{m1} , v_{m1} , v_{m1} , $f_1(v_{m1})$ and $f_1(v_{m1})$ (Fig. 17a). In this case

$$d_1 = \frac{v_{m1}}{v_{m1}}. \quad (1-201)$$

Under the made assumptions the mechanism of the formation of FAKh of multistage amplifier considerably is complicated. Let us consider the process of formation of FAKh on the block diagram in Fig. 16c. For simplification in the analysis upon the saturation of the i cascade/stage on the formation of FAKh we will consider the

output effect only of one of previous (i-1) cascade/stage.

Let us assume that with E_1 the latter/last (n-th) cascade/stage is saturated. Then

$$\mathcal{G}_1 = F(E_1) = f_n(v_{nn})k_n + f_{n-1}(v_{n-1})k_{n-1}.$$

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In this case there can be two cases:

1. The inequality

$$\frac{v_{nn}}{K_{nn-1}} < v_{nn-1} \quad (1-202)$$

is fulfilled. Φ In this case (n-1) cascade/stage works in the linear conditions and its output effect can be expressed through its initial transmission factor K_{nn-1} .

2. Inequality

$$\frac{v_{nn}}{K_{nn-1}} > v_{nn-1} \quad (1-203)$$

is fulfilled. Φ In this case (n-1) cascade/stage works in the nonlinear mode/conditions, and, if analytical writing $f_{n-1}(v_{n-1})$ is unknown then it cannot be expressed $f_{n-1}(v)$ through K_{nn-1} .

Since $f_{n-1}(v) = v_n$,

$$\mathcal{G}_1 = k_n f_n(v_{nn}) + k_{n-1} v_{nn}.$$

two terms 3)

$$\left. \begin{aligned} k_{n-1} &= \frac{k_n / n (v_{nn}) (0_{n-1} - 1)}{f_{n-1} (v_{nn-1}) - v_{nn}}; \\ &\dots \dots \dots \\ k_{n-m} &= \frac{\left[\sum_{i=0}^{m-1} k_{n-i} / n-i (v_{nn-i}) \right] (0_{n-m} - 1)}{f_{n-m} (v_{nn-m}) - v_{nn-m+1}}. \end{aligned} \right\} \quad (1-205)$$

In the case in question the objective parameter of amplifier will coincide with the point of FAKh only at the isolated points as with the linear cascades/stages. However, intermediate points will deviate in the smaller measure.

For obtaining precise FAKh of amplifier nonlinear cascades/stages must have completely specific characteristics, whose form can be determined from the working conditions for the successive of identical and different parallel pairs.

Analysis FU of the parallel type.

The requirements, presented to amplifier stages and corrective elements/cells, in parallel type FU depend on the circuit solution of supplying the effect on inputs of cascades/stages. Effect can enter directly the inputs of all cascades/stages (Fig. 16b); through the divider (Fig. 16d and f), also, through the corrective elements/cells (Fig. 16h).

FU without the corrective elements/cells (Fig. 16b). Diagram in Fig. 16b is a special case of diagram in Fig. 16h, when to the cascades/stages are not given cutoff voltages and $k_i = 1$.

Amplitude characteristic of n-cascade FU is described by the expression

$$F(E) = \sum_{i=1}^n f_i(v). \quad (1-206)$$

Initial transmission factor of amplifier with the work in the linear conditions

$$K_0 = \sum_{i=1}^n K_{0i}. \quad (1-207)$$

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For obtaining precise FAKh the cascades/stages must be identical, nonlinear and have AKh, described by the expression

$$f_i(v_i) = f(E) = \frac{F(E)}{n}, \quad (1-208)$$

where $F(E)$ is described by expressions (1-15) and (1-19).

The slope/transconductance of FAKh of the amplifier

$$\alpha_0 = \sum_{i=1}^n \alpha_i. \quad (1-209)$$

where α_i — slope/transconductance of FAKh of the i cascade/stage.

If cascades/stages are identical, the slope/transconductance of FAKh of amplifier is n times more than the slope/transconductance of FAKh of cascade/stage, which is the advantage of diagram.

In the diagram in question the dynamic range of FAKh of amplifier is equal to the dynamic range of the nonlinear section of the characteristic of amplifier, i.e., $D=d$, what is a deficiency/lack in this diagram.

Diagrams with the divider and the corrective elements/cells, in which it is possible to carry out a successive work of amplifier stages, possess considerably greater possibilities.

FU with the different linear cascades/stages, which work in the mode/conditions of amplification-limitation and without the cutoff voltages. It is obvious that parallel type FU with the different cascades/stages it is possible to fulfill on one of the diagrams (Fig. 16d, f, h).

Analyzing diagram with the divider at the input (Fig. 16d), we assume that the cascades/stages have the amplitude characteristics, depicted in Fig. 17a, and different values $k_{\text{н}}$ and $\mu_{\text{н}}$. Let us agree

to designate the coefficients of the divisions of divider through k_i . Coefficient of the division of input effect for the i cascade/stage with $k_1=1$ according to Fig. 16d

$$k_i = \frac{\sum_{m=1}^n R_m}{\sum_{m=1}^n R_m} \quad (1-210)$$

Let us consider the work of amplifier. Let us assume that when $E_1 = v_{n1}$ the 1st cascade/stage (Fig. 18b) is saturated. Then effect at the output of the amplifier

$$\vartheta_1 = E_1 (k_1 K_1 + k_1 k_2 K_2 + \dots + K_n \prod_{i=1}^n k_i).$$

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In this case effect at the input of the 2nd cascade/stage

$$v_2 = E_1 k_2 = v_{n1} k_2.$$

With $E_2 = E_1$ is saturated the 2nd cascade/stage and

$$\vartheta_2 = E_1 [k_1 K_1 + d_2 (k_1 k_2 K_2 + k_1 k_2 k_3 K_3 + \dots + K_n \prod_{i=1}^n k_i)], \quad (1-211)$$

where $d_2 = \frac{E_2}{E_1} = \frac{v_{n2}}{v_{n1}} = \frac{v_{n2}}{v_{n1} k_2}$ — working dynamic range of 2nd cascade/stage, which in Fig. 18b is designated by numeral 2;

with $E_3 = E_1 d_2 d_3$ is saturated the 3rd cascade/stage and

$$\vartheta_3 = E_1 [k_1 K_1 + d_2 k_1 k_2 K_2 + d_2 d_3 (k_1 k_2 k_3 K_3 + \dots + K_n \prod_{i=1}^n k_i)],$$

where

$$d_2 = \frac{v_{K2}}{v_{K1} d_2 k_2 K_2} = \frac{v_{K2}}{E_1 d_2 k_2 K_2};$$

when $E_{n-1} = E_1 \prod_{i=1}^{n-1} d_i$ is saturated (n-1) cascade/stage and then

$$\mathcal{D}_{n-1} = E_1 (k_1 K_1 + k_1 k_2 d_2 K_2 + \dots + K_n \prod_{i=1}^{n-1} d_i \prod_{i=1}^n k_i); \quad (1-212)$$

when $\mathcal{D}_n = E_1 \prod_{i=1}^n d_i$ is saturated the n cascade/stage and

$$\mathcal{D}_n = E_1 (k_1 K_1 + k_1 k_2 d_2 K_2 + \dots + K_n \prod_{i=1}^{n-1} d_i \prod_{i=1}^n k_i), \quad (1-213)$$

where

$$d_n = \frac{v_{Kn}}{E_1 \prod_{i=1}^{n-1} d_i \prod_{i=1}^n k_i}. \quad (1-214)$$

Taking into account reaction only of one subsequent in parallel connected cascade/stage, and also condition (1-142), we find expression for the coefficients of division in the implementation of LAKh:

$$\left. \begin{aligned} k_2 &= \frac{\sigma \ln d_2}{E_1 (d_2 - 1) K_2}; \quad k_3 = \frac{\sigma \ln d_3}{E_1 (d_3 - 1) d_2 k_2 K_3}; \\ &\dots \dots \dots \\ k_n &= \frac{\sigma \ln d_n}{E_1 (d_n - 1) \prod_{i=1}^{n-1} d_i \prod_{i=1}^{n-1} k_i} \end{aligned} \right\} \quad (1-215)$$

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If objective parameters of UK are known, then the coefficients of division can be calculated from the formula

$$k_i = \frac{v_{ni} R_{i-1}}{R_i v_{n(i-1)} d_i} \quad (1-216)$$

and it is possible to determine slope/transconductance of LAKh from the parameters of the characteristics of amplifier stages

$$a_1 = \frac{v_{n2}(d_1 - 1)}{d_1 \ln d_1}; a_2 = \frac{v_{n3}(d_2 - 1)}{d_2 \ln d_2}; \dots; a_n = \frac{v_{nn}(d_{n-1})}{d_n \ln d_n}. \quad (1-217)$$

Equations (1-217) are transcendental and can be solved graphically or by the method of iterations.

If condition $d_i \gg 1$ is satisfied then slope/transconductance tentatively can be calculated from the approximation formulas

$$a_1 = \frac{v_{n2}}{\ln d_1}; a_2 = \frac{v_{n3}}{\ln d_2}; \dots; a_n = \frac{v_{nn}}{\ln d_n}. \quad (1-218)$$

Expressions (1-217) and (1-218) can be used for the check calculations of finished diagrams.

Analogously it is possible to show that for the diagram in Fig. 16h when $E_1 = \bar{v}_n$ and $k=1$ is correct the expression for the transmission factor of the i corrective element/cell, connected at the input

$$k_i = \frac{v_{ni}}{E_1 R_i \prod_{m=1}^i d_m}, \quad (1-219)$$

and also expression (1-217) and (1-218).

For the realization of SAKh in the diagram in Fig. 16h taking into account (1-148) and reacting only the saturated cascades/stages we obtain

$$\left. \begin{aligned} k_1 &= \frac{R_1(\theta_1 - 1)}{d_1 R_1}; \quad k_2 = \frac{(\theta_2 - 1)(k_1 R_1 + d_1 k_1 k_2 R_2)}{d_2 d_1 k_2 R_2}; \\ &\dots \dots \dots \\ k_n &= \frac{(\theta_n - 1)(k_1 R_1 + d_1 k_1 k_2 R_2 + \dots + k_{n-1} \prod_{i=1}^{n-1} d_i \prod_{j=1}^{n-1} k_j)}{R_n \prod_{i=1}^{n-1} k_i \prod_{j=1}^n d_j} \end{aligned} \right\} \quad (1-220)$$

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Analogously for the diagram in Fig. 16h taking into account only the reaction of the saturated cascades/stages, we obtain

$$\left. \begin{aligned} k_1 &= \frac{k_1 R_1 (\theta_1 - 1)}{d_1 R_1}; \\ k_2 &= \frac{(\theta_2 - 1)(k_1 R_1 + d_1 k_1 k_2 R_2)}{d_2 d_1 k_2 R_2}; \\ &\dots \dots \dots \\ k_n &= \frac{(\theta_n - 1)(k_1 R_1 + d_1 k_1 k_2 R_2 + d_2 d_1 k_2 k_3 R_3 + \dots + k_{n-1} R_n \prod_{i=1}^{n-1} d_i)}{k_1 R_1 + d_1 k_1 k_2 R_2 + d_2 d_1 k_2 k_3 R_3 + \dots + k_{n-1} R_{n-1} \prod_{i=1}^{n-1} d_i} \end{aligned} \right\} \quad (1-221)$$

On the basis of analysis it is possible to do the conclusion that upon the parallel connection of the different cascades/stages, which work in the mode/conditions of linear amplification-limitation and without the cutoff voltages, cannot be realized FAKh with $b = \text{var}(\uparrow)$ in ShDD with the successive work of cascades/stages.

However, FAKh with $b = \text{var}(\uparrow)$ it is possible to realize with the successive work of cascades/stages in the amplifier with the corrective elements/cells, connected at inputs or outputs of cascades/stages (1, 2, ... in Fig. 16h). Cascades/stages alternately enter into the operational conditions during the supplying of the corresponding cutoff voltages on the corrective elements/cells.

The in principle cutoff voltages can be supplied to amplifier stages. However, in the latter case further difficulties in the formation of FAKh appear, since the differential transmission factor of cascade/stage on leaving from the cutoff mode/conditions vanishes.

FAKh in the amplifier, assembled on the diagram in Fig. 16h, can be formed with the work of amplifier stages in the linear conditions in entire dynamic range of FAKh of amplifier; in the mode/conditions of linear amplification-limitation; in the nonlinear mode/conditions.

The second and third modes/conditions of cascades/stages are suitable for formation of FAKh in the amplifier, assembled on any diagram in Fig. 16.

FU with the different cascades/stages and the corrective elements/cells (Fig. 16h).

1. Cascades/stages work in linear conditions.

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For guaranteeing the successive work of cascades/stages on the input corrective (untying) elements/cells the cutoff voltages

$$U_{s1}=0; U_{s2}=E_1; U_{s2}=E_2=E_1d_1, \dots, U_{sn}=E_{n-1}=E_1 \prod_{i=1}^{n-1} d_i \quad (1-222)$$

are given. If cutoff voltages are supplied to the output untying (corrective) elements/cells, then their values are calculated from the formula

$$U_{st}'' = K_1 U_{st}' = K_1 E_{t-1} \prod_{m=1}^{t-1} d_m. \quad (1-223)$$

Let us consider work of FU. With $E \leq E_1$, the output of amplifier the effect enters only from the output of the first cascade/stage and the transmission factor of amplifier is equal to the transmission factor of the first amplifier element/cell

$$K_0 = k'_1 k''_1 K_{n1} = K_1, \quad (1-224)$$

where k'_1 , k''_1 , and K_{n1} — respectively the transmission factors of the input and output corrective elements/cells and amplifier stage;

$K_1 = k'_1 k''_1 K_{n1}$ — transmission factor of the first amplifier element/cell, which consists of series-connected UK and corrective elements/cells.

With $E_1 \vartheta_1 = K_1 E_1$;

with $E \geq E_1$ together with the first cascade/stage works the second; then (Fig. 21)

$$\vartheta = F(E) = E(K_1 + K_2) - E_1 K_2;$$

where $E_1 = E_1 d_1$,

$$\vartheta_1 = E_1 [d_1 (K_1 + K_2) - K_2]; \quad (1-225)$$

with $E_2 = E_1 d_1 d_2$,

$$\vartheta_2 = E_1 [d_1 d_2 (K_1 + K_2 + K_3) - K_2 - d_2 K_3];$$

when $E_n = E_{n-1} d_n = E_1 \prod_{i=1}^n d_i$

$$\vartheta_n = E_1 \left[\left(\prod_{i=1}^n d_i \right) \sum_{i=1}^n K_i - K_2 - d_2 K_3 - d_2 d_3 K_4 - K_n \prod_{i=2}^{n-1} d_i \right].$$

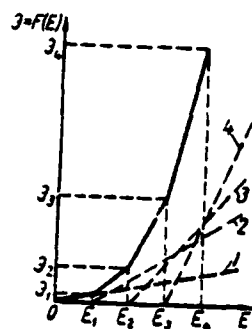


Fig. 21. FAKh with $b = \text{var}(\uparrow)$ of the amplifier, assembled on the block diagram Fig. 16h.

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Dynamic ranges of FAKh on the input effect and output effect respectively are equal to:

$$D = \prod_{i=1}^n d_i; \quad (1-226)$$

$$D_{\text{max}} = \prod_{i=1}^n d_{\text{max},i}. \quad (1-227)$$

Using conditions (1-142), (1-147) and (1-148), we find formulas for determining the transmission factors of amplifier elements/cells in the implementation of LAKh:

$$\left. \begin{aligned} K_1 &= \frac{\sigma \ln d_1}{E_1(d_1 - 1)} - K_1; \quad K_2 = \frac{\sigma \ln d_2}{E_2 d_2(d_2 - 1)} - (K_1 + K_2); \\ &\dots \dots \dots \\ K_n &= \frac{\sigma \ln d_n}{E_n(d_n - 1) \prod_{i=1}^{n-1} d_i} - \sum_{i=1}^{n-1} K_i; \end{aligned} \right\} \quad (1-228)$$

in the implementation of SAKh and PAKh:

$$\left. \begin{aligned} K_1 &= \frac{K_1(d_1-1)}{d_1-1}; K_2 = \\ &= \frac{d_2[d_1(K_1+K_2)-K_1]-d_1d_2(K_1+K_2)+K_1}{d_2(d_2-1)}; \\ &\dots\dots\dots \\ K_n &= \frac{d_n \left[\prod_{i=2}^{n-1} d_i \sum_{i=1}^n K_i - A \right] + A - \prod_{i=2}^n d_i \sum_{i=1}^{n-1} K_i}{(d_n-1) \prod_{i=2}^{n-1} d_i} \end{aligned} \right\} \quad (1-229)$$

where $A = K_1 + d_1K_2 + d_1d_2K_3 + \dots + K_{n-1} \prod_{i=2}^{n-1} d_i$; $0_n = d_n^2$ in the implementation of SAKh; $0_n = M^{2E_1(d_n-1) \prod_{i=2}^{n-1} d_i}$ in the implementation of PAKh.

The advantage of the mode of operation of cascades/stages examined is the fact that FAKh with $b = \text{var}(\hat{f})$ can be formed with the transmission factors of the amplifier elements/cells, equal to one, i.e., with satisfaction of the condition

$$K_1 = K_2 = \dots = K_n = 1. \quad (1-230)$$

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However, this mode/conditions has the deficiency that FAKh can be formed in a comparatively small dynamic range of the first cascade/stage.

From the deficiency/lack indicated is free the mode/conditions of linear amplification-limitation.

2. Amplifier elements/cells work in mode/conditions of linear amplification-limitation.

The mode/conditions of the limitation of amplifier element/cell can be realized directly in amplifier stage either in the untying element/cell, connected to input or output of amplifier stage (Fig. 16h). If it is necessary to sharply bound signal level, from limiting element/cell should be included at the output of amplifier stage, since with large levels of signal an effect of limitation to carry out more easily.

In each special case the connection point of limiting element/cell is determined by the circuit solution, the type of amplifier element/cell and by form of FAKh.

Let us consider the strictly successive work of amplifier elements/cells in the mode/conditions of amplification-limitation on the assumption that the limiting elements/cells are connected to the inputs of amplifier stages and on them are given the cutoff voltages, whose values are selected from condition (1-222).

The output effect of amplifier with the work of the i

cascade/stage in the linear conditions is changed according to the law

$$\mathcal{D}_i = \sum_{m=1}^{i-1} f_m(v_m) + (E_i - E_{i-1}) K_i. \quad (1-231)$$

The output effects of amplifier, which correspond to levels $E_1, E_2, E_3, \dots, E_n$, with which they enter into the mode/conditions of limitation 1, 2, ..., the n cascade/stage, are equal to

$$\begin{aligned} \mathcal{D}_1 &= E_1 K_1; \\ \mathcal{D}_2 &= E_1 [K_1 + K_2 (d_2 - 1)] = \mathcal{D}_1 + E_1 K_2 (d_2 - 1); \\ \mathcal{D}_3 &= E_1 [K_1 + K_2 (d_2 - 1) + K_3 d_2 (d_2 - 1)] = \mathcal{D}_2 + E_1 K_3 d_2 (d_2 - 1); \\ &\dots \dots \dots \\ \mathcal{D}_n &= E_1 [K_1 + K_2 (d_2 - 1) + K_3 d_2 (d_2 - 1) + \dots + K_n (d_n - 1) \prod_{i=1}^{n-1} d_i]. \end{aligned}$$

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Taking into account conditions (1-140), (1-142), (1-147) and (1-148), we obtain, that for the realization of linear characteristic the transmission factors of all amplifier elements/cells must be equal, i.e.

$$K_1 = K_2 = \dots = K_n = K; \quad (1-232)$$

for the realization of LAKh:

$$K_1 = \frac{e \ln d_1}{E_1 (d_1 - 1)}; K_2 = \frac{e \ln d_2}{E_1 d_1 (d_2 - 1)}; \dots; K_n = \frac{e \ln d_n}{E_1 (d_n - 1) \prod_{i=1}^{n-1} d_i}; \quad (1-233)$$

for the realization of SAKh and PAKh:

$$K_1 = \frac{K_1(\theta_1 - 1)}{d_1 - 1}; K_2 = \frac{[K_1 + K_2(d_1 - 1)](\theta_2 - 1)}{d_2(d_1 - 1)}; \dots;$$

$$K_n = \frac{[K_1 + K_2(d_1 - 1) + K_3 d_1(d_2 - 1) + \dots + K_{n-1}(d_{n-1} - 1)] \prod_{i=1}^{n-1} d_i (\theta_n - 1)}{(d_n - 1) \prod_{i=1}^{n-1} d_i} \quad (1-234)$$

In the case in question in the implementation of FAKh with $b = \text{var}(\uparrow)$ is implemented inequality

$$K_1 < K_2 < K_3 < \dots < K_n. \quad (1-235)$$

in the implementation of FAKh with $b = \text{var}(\uparrow)$ is implemented the inequality

$$K_1 > K_2 > \dots > K_{n-1} > K_n. \quad (1-236)$$

which can be realized at the given values of transmission factors UK with the help of the selection of the transmission factors of the corrective elements/cells.

By the advantage of the method examined is the possibility of obtaining FAKh in the broad dynamic band with the input effect;

deficiency/lack - smaller slope/transconductance of FAKh, than with the work of cascades/stages in the linear conditions with the same transmission factors.

According to the obtained recurrent formulas (1-228), (1-229), (1-232), and (1-233), we can calculate the required transmission factors of cascades/stages, if are assigned slope/transconductance of FAKh (or its law), the input effect E_1 , under which it must begin, and the working dynamic ranges of the cascades/stages on the input effect, whose value is determined by the assigned accuracy of realization of FAKh.

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In practice can arise the inverse problem, when the transmission factors of amplifier elements/cells are known and it is necessary to determine the working dynamic ranges of cascades/stages d_1 on the basis of the assigned to accuracy of realization of FAKh. In the case of amplifier with LAKh for determination d_1 from the assigned accuracy of the realization of characteristic it is possible to use (1-177), (1-182) and (1-183).

For the remaining types of FAKh it is possible to obtain analogous formulas. After determination d_1 from formulas (1-222) or (1-223) the cutoff voltages, supplied are designed for the corrective elements/cells.

Parallel type of FU with the nonlinear cascades/stages.

It is obvious that during the use of linear cascades/stages only the isolated points of objective parameter coincide with precise FAKh. The points of the coincidence of characteristics are greater, the more the cascades/stages participates in the formation of FAKh. However, it can seem that the assigned accuracy of FAKh can be obtained with an inadmissibly large number of cascades/stages. In this case for the formation of FAKh it is necessary to apply nonlinear cascades/stages.

For the formation of more precise FAKh should be taken the nonlinear cascades/stages, which have the same character of a change in the differential transmission factor (increase, decrease).

In the most general case, when AKh cascades/stages cannot be described mathematically, the transmission factors of nonlinear cascades/stages it is necessary to calculate from formulas (1-204), (1-205), (1-228), (1-229), (1-233) and (1-234) for the levels of input effect, which correspond to the moments/torques of the entrance of cascades/stages into the mode/conditions of limitation (saturation). For the n cascade/stage during the supplying of cutoff voltages on the corrective elements/cells this level

$$v_{nk} = v_n - v_{n-1} = E_1 (d_n - 1) \prod_{i=1}^{n-1} d_i. \quad (1-237)$$

For obtaining FAKh of the high accuracy AKh of cascades/stages must be completely specific. The required form of the characteristic of nonlinear cascade/stage can be determined, on the basis of the principle of the successive operation of the parallel pair of cascades/stages. In this case the i nonlinear cascade/stage taking into account output effects $(i+1)$, $(i+2)$, ... cascades/stages must form/shape AKh, suitable for the successive work of cascades/stages and nonlinear mode/conditions.

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If nonlinear cascades/stages are identical and work without the cutoff voltages, AKh cascades/stages with the strictly successive work just as in consecutive type FU, they must be described by equations (1-190) and (1-191), in which it is necessary to substitute $F(E)$ for f_1 and K_0 instead of K_1 . The initial transmission factor of cascade/stage is determined from expression (1-192).

For the realization of FAKh in the diagrams in Fig. 16d, f must be implemented the equality

$$k_1 = \frac{1}{2}, \quad (1-238)$$

where for the amplifier with LAKh $d = K_0$, for the amplifier with SAKh

when $b = \text{var}(\downarrow) - d = K_n^{\frac{n}{n-1}}$.

Then the initial transmission factor of the amplifier, made on the diagram in Fig. 16d,

$$K_0 = k_1 K_1 + k_1 k_2 K_2 + k_1 k_2 k_3 K_3 + \dots + K_n \prod_{i=1}^n k_i. \quad (1-239)$$

Usually $k_1=1$. The values of the resistors/resistances of divider are calculated, on the basis of formulas (1-210) and (1-238).

For the realization of LAKh in the diagram in Fig. 16h the equalities

$$\left. \begin{aligned} K_{n1} &= d_1; K_{n2} = \frac{K_{n1}}{d_1} = 1; K_{n3} = \frac{K_{n2}}{d_2} = \frac{1}{d}, \dots, \\ K_{nm} &= \frac{K_{n1}}{d_1^{m-1}} = \frac{1}{d^{m-1}}, \end{aligned} \right\} \quad (1-240)$$

must be implemented since

$$d_1 = d_2 = \dots = d_n = d. \quad (1-241)$$

In this case

$$v_{n2} = v_{n1} d_1; v_{n3} = v_{n1} d_1^2, \dots, v_{nm} = v_{n1} d_1^{m-1}.$$

General/common/total initial transmission factor of the amplifier

$$K_0 = \sum_{i=1}^n K_{ni} = \sum_{i=1}^n \frac{K_{n1}}{d^{i-1}}. \quad (1-242)$$

In the implementation of SAKh in the diagram in Fig. 16h must be satisfied the condition

$$d_1 = K_n^{\frac{n}{n-1}}.$$

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Then

$$K_{n2} = \frac{K_{n1}}{d_1}; K_{n3} = \frac{K_{n2}}{d_2} = \frac{K_{n1}}{d_1 d_2}, \dots, K_{nn} = \frac{K_{n1}}{\prod_{i=1}^{n-1} d_i}; \quad (1-243)$$

$$v_{n2} = v_{n1} d_1, v_{n3} = v_{n1} d_1 d_2, \dots, v_{nn} = v_{n1} \prod_{i=1}^{n-1} d_i. \quad (1-244)$$

If to output corrective elements/cells in the diagram in Fig. 16h cutoff voltages are given, then the amplitude characteristics of the nonlinear cascades/stages, which strictly alternately work in the functional modes/conditions, must be described by expressions (1-20), (1-24) and (1-200). Furthermore, must be satisfied condition (1-76).

Cutoff voltages must be equal to:

$$U_{et} = f_t(v_{nt}). \quad (1-245)$$

For the fulfillment of equality (1-200) it is necessary in each cascade/stage to bound output effect at the level $f_t(v_{nt})$, using for this the output corrective elements/cells.

In any event in the implementation of FAKh on the diagram in Fig. 16h it is necessary to satisfy the following condition: the

initial differential transmission factor of n cascade/stage b_{nn} must be equal to final differential transmission factor k_{nn-1} previous (n-1) cascade/stage, i.e.

$$b_{nn} = b_{nn-1}. \quad (1-246)$$

Condition (1-246) is satisfied, if the equality

$$K_{nn} = b_{nn-1}, \quad (1-247)$$

is satisfied since

$$b_{nn} = K_{nn}.$$

Thus, with the successive work of nonlinear cascades/stages in the diagram in Fig. 16h initial transmission factor of the n cascade/stage must be equal to differential transmission factor of (n-1) cascade/stage with its work at the end of the functional mode/conditions.

The value of coefficients can be calculated from the formulas for b, given in Table 1, when $x = x_{ni} = d_i$, where d_i — dynamic range of FAKh of the i cascade/stage.

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§6. Method of the repeated use of a functional amplifier element/cell.

Realization of FAKh in ShDD on one functional amplifier

element/cell (FUE) is limited to the dynamic range of the functional section either of the passage characteristic of amplifier instrument, if the function of amplification and formation of FAKh perform the nonlinear amplifier instrument (UP), or the nonlinear element/cell, which forms FAKh. FAKh it is possible to obtain in ShDD according to the method of repeated use of one FUE.

Let us consider essence and possible methods of the application of this method. Let us assume that FAKh of amplifier element/cell begins and is finished under input influences v_n and v_m . Then

$$d = \frac{v_n}{v_m}.$$

Let us assume that the effect at the input of FUE achieved value v_n . For a second time and, consequently, also it is repeated, it is possible to use FUE, if when $v = v_n$ it is abrupt: 1) to widen the dynamic of the amplifier element/cell; 2) to change the level of effect at the input of FUE; 3) to decrease the differential transmission factor of amplifier element/cell.

Some possible block diagrams of the application of the method of obtaining FAKh in question are given on Fig. 22. As we see, signals can enter the input of FUE directly (Fig. 22a) or through the adjustable cascades/stages (RK, Fig. 22b). In both cases in parallel FUE and RK are connected the adjusters (RU), which on the specific levels of input effect develop controlling voltages/stresses U_{pi} .

which enter on FUE or on RK,.

The mechanism of obtaining FAKh let us consider on the block diagram of the amplifier Fig. 22b, which has a series/row of advantages and is more spread. With a change in the effect at the input of FU from $E_1 = E_n$ to $E_2 = E_d$ the effect at the input of FUE varies from v_n to v_m and FUE works in the functional mode/conditions.

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On the level of effect E_1 adjuster PV_n operates/wears and develops controlling voltage/stress U_{pn} , which enters adjustable cascade/stage PK_n . The transmission factor of cascade/stage PK_n is reduced in m of times. If $m=d$, then effect at the input of FEU is equal to v_n .

With further increase of the input effect of the amplifier from E_1 to $E_2 = E_d$, FEU works in functional mode/conditions. With E_1 effect at the input of FEU reaches v_n . At this moment PV_{n-1} operates/wears transmission factors PK_{n-1} is reduced m times.

Each time, when the input effect of the amplifier reaches value $E_1 = E_d$, with which the level of effect at the input of FEU is equal to v_n operates/wears PV_1 and is reduced transmission factor PK_1 to m of times.

If there are n adjustable cascades/stages, then FEU is used $n+1$ times. As a result discontinuous FAKh is obtained. With the depth of the adjustment of each cascade/stage $m=d$ the dynamic range of discontinuous FAKh is determined by expression (1-59).

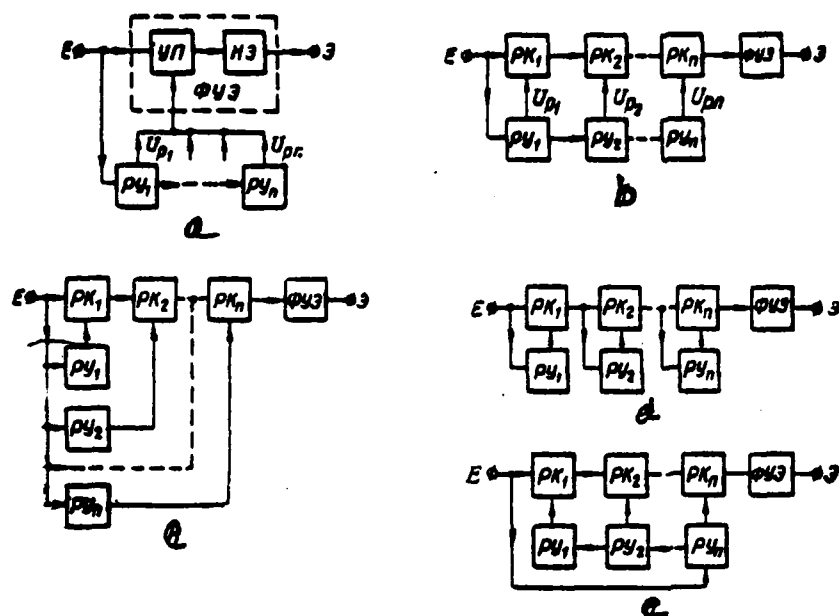


Fig. 22. Block diagrams of amplifiers with discontinuous FAKh.

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Gain control in PK_1 is equivalent to an abrupt change of the differential transmission factor FU as a whole (Fig. 22a).

In the case of the block diagram Fig. 22a controlling voltage/stress must change either differential transmission factor of UP or dynamic range of the nonlinear section of passage characteristic UP (in the absence of the nonlinear element/cell (NE), which forms FAKh). The dynamic range of passage characteristic can be

expanded by an increase in the depth of negative feedback [56]. As can be seen from Fig. 22 FU can be made with the adjustment forward (Fig. 22c, d) and back/ago (Fig. 22e) upon the inclusion/connection RU in parallel (Fig. 22c) and consecutively/serially (Fig. 22e).

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Chapter 2.

FUNCTIONAL AMPLIFIERS WITH NONLINEAR ELEMENTS.

§ 1. Functional amplifiers as a nonlinear quadrupole.

Amplifier stage can be the simplest functional amplifier, basic component elements of which are amplifier instrument, load and power supply. The form of the amplitude characteristic of amplifier is determined by the parameters of each of the elements/cells. During the appropriate selection of power supply the effect of the latter on the form of FAKh it is possible not to consider. Then FAKh in amplifier stage can be realized due to the nonlinear properties of UP or the load or due to nonlinearity of both elements/cells.

Vacuum tubes and transistors are basic amplifier instruments in the electronic amplifiers. If electron tube with the sufficiently low voltage inputs with some assumptions can be considered linear instrument, then transistor according to the character of the dependence of currents on the stresses/voltages is in principle nonlinear device. With a strict approach electron tube also should be considered nonlinear device even with the low signals.

For the formation of FAKh as the nonlinear elements/cells are applied vacuum and semiconductor diodes, transistors, and also their different combinations with the linear resistance.

In the general case of one-stage FU it is possible to represent in the form of the nonlinear quadrupole (Fig. 23), the characteristic parameters and load of which depend on the level of voltage input U_1 . We analyze amplifier and determine how loads must change and the parameters UP for the realization of required FAKh.

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Since as UP in essence vacuum tubes and transistors are applied, for the analysis we select the system of Y-parameters [22, 87], which makes it possible to utilize the detailed theory and procedure of calculation of vacuum-tube circuits for the transistor amplifiers, as a result possible to obtain the overall dependences, suitable both for the electron-tube and transistor functional amplifiers; the recording of fundamental principles for different amplifiers in this system of the parameters simplest; all parameters are measured comparatively easily and sufficiently simply; the parameters clearly depend on those factors, which substantially affect the qualitative indices of amplifier; in the transistor amplifiers Y-parameter system relates to the basic most widely used connection of transistor - with

the general/common/total emitter.

In Y-parameter system for independent variables the stresses/voltages accept and the equation of network are written/recorded in the form

$$\left. \begin{aligned} I_1 &= Y_{11}\dot{U}_1 + Y_{12}\dot{U}_2 \\ I_2 &= Y_{21}\dot{U}_1 + Y_{22}\dot{U}_2 \end{aligned} \right\} (2-4)$$

Being based on works [22, 56], let us examine Y-parameters of tube and transistor and will set between them interconnection and analogy.

The equivalent schematics of tube and high-frequency diffusion transistor [22] are given in Fig. 24. With some assumptions [22] the diagram in Fig. 24b is suitable for the drift transistors. Subsequently we will use only one term "high-frequency" transistors, implying diffusion and drift transistors.

Diagram in Fig. 24b is one of the known versions of equivalent diagrams, but it has a number of the deficiencies/lacks, basic from which is nonconformity not to one diagram, which escape/ensues from the theory of quadrupoles. The entering it values do not yield to direct measurement, but the controlled current is expressed as the unknown stress/voltage on the emitter junction.

For eliminating the deficiencies/lacks the diagram must be

converted into the equivalent schematic of quadrupole in Y-parameter system. This conversion is carried out in work [22]. In this case the digital indexing of Y-parameters is replaced to the designations, which open the physical sense of Y-parameters,.

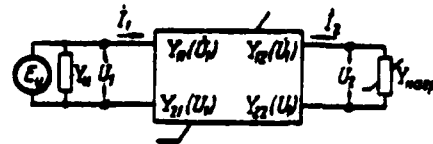


Fig. 23. Functional amplifier as nonlinear quadrupole.

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After the replacement of the designations of the equation of currents (2-1) for the transistor they take the following form:

$$\left. \begin{aligned} I_1 &= Y\dot{U}_0 - Y_{00p}\dot{U}_2; \\ I_2 &= S\dot{U}_0 + Y_1\dot{U}_2. \end{aligned} \right\} \quad (2-2)$$

where $Y=Y_{11}$ - input admittance of transistor; $Y_{00p}=Y_{12}$ - conductivity of feedback; $S=Y_{21}$ - slope/transconductance of the passage characteristic of transistor; $Y_1=Y_{22}$ - the internal conductance of equivalent generator.

In the range of the frequencies the conductivities have complex character and are determined from the formulas, given in appendix 1. At the lowest frequencies all conductivities active, do not depend on frequency. They can be expressed as the usual differential parameters:

$$\left. \begin{aligned} Y_{11-0} = g &= \frac{\partial i_1}{\partial U_0} \Big|_{U_2=\text{const}}; & S_{12-0} = S_0 &= \frac{\partial i_2}{\partial U_0} \Big|_{U_2=\text{const}}; \\ Y_{00p-0} = g_{00p} &= -\frac{\partial i_1}{\partial U_2} \Big|_{U_0=\text{const}}; & Y_{22-0} = \frac{1}{R_1} = g_1 &= -\frac{\partial i_2}{\partial U_2} \Big|_{U_0=\text{const}}. \end{aligned} \right\} \quad (2-3)$$

U-shaped equivalent circuit for the tube and the transistor in Y-parameter system are given in Fig. 25.

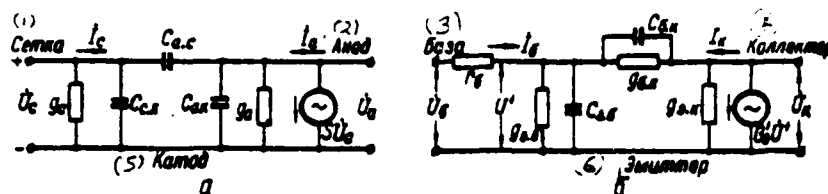


Fig. 24. Equivalent diagrams: a) vacuum tube; b) high-frequency transistor.

Key: (1). Grid. (2). Anode. (3). Base. (4). Collector/receptacle. (5). Cathode. (6). Emitter.

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Comparing the equivalent systems of amplifier instruments, and also equation (2-2) with the equations of the grid and anode currents of tube [56]

$$I_g = (Y_c + Y_{a, a}) \dot{U}_c - Y_{a, g} \dot{U}_a; \quad I_a = (S - Y_{a, a}) \dot{U}_c + (Y_g + Y_{a, g}) \dot{U}_a. \quad (2-4)$$

it is easy to set the analogy between the transistors and the tubes: common cathode - general/common/total emitter; common grid - general/common/total base; the general/common/total anode - common collector/receptacle (Fig. 26).

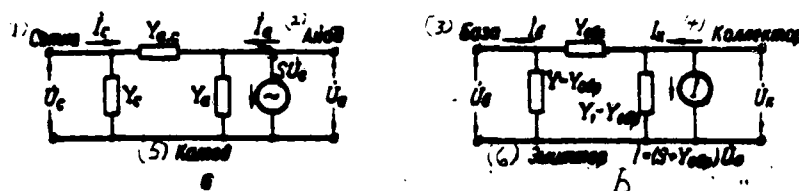


Fig. 25. U-shaped equivalent circuits and Y-parameter system: a) for the tube; b) for the transistor.

Key: (1). Grid. (2). Anode. (3). Base. (4). Collector/receptacle. (5). Cathode. (6). Emitter.

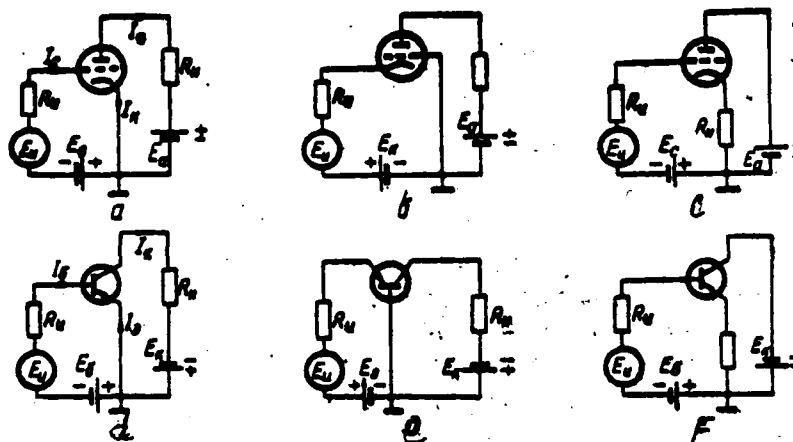


Fig. 26. Connections of tube and transistor of type n-p-n with general/common/total ones: a) by cathode; b) by grid; c) by anode; d) by emitter; e) by base; f) by collector/receptacle.

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According to the theory of quadrupoles for the diagram with the common cathode we find;

the factor of amplification of cascade/stage in the stress/voltage

$$K = \frac{U_a}{U_o} = - \frac{S - Y_{a.o}}{Y_{a.o} + Y_a + Y_{narp}}; \quad (2-5)$$

input admittance of the cascade/stage

$$Y_{ax} = \frac{i_o}{U_o} = Y_o + (1 - K) Y_{a.o}; \quad (2-6)$$

the output admittance of the cascade/stage

$$Y_{max} = \frac{i_a}{U_a} = Y_a + Y_{a.o} + \frac{(S - Y_{a.o}) Y_{a.o}}{Y_n + Y_o + Y_{a.o}}, \quad (2-7)$$

where Y_{narp} - load admittance; Y_n - the internal conductance of the source of signal (or the previous cascade/stage).

Replacing for the conductivity of tube Y_o , $Y_{a.o}$, S , Y_a in expressions (2.5), (2.6), (2.7) respectively by values

$Y - Y_{oop}$, Y_{oop} , $S + Y_{oop}$, $Y_i - Y_{oop}$, we obtain expressions for determining the parameters of common-emitter connection

$$K = \frac{U_n}{U_o} = \frac{S}{Y_i + Y_{narp}}; \quad (2-8)$$

$$Y_{ax} = \frac{i_o}{U_o} = Y - K Y_{oop}; \quad (2-9)$$

$$Y_{max} = \frac{i_a}{U_n} = Y_i + \frac{S Y_{oop}}{Y_n + Y}. \quad (2-10)$$

Analogously it is possible to obtain expressions for the parameters of cascade/stage at the inclusion/connection of UP on the diagram with the general/common/total ones by grid and the base, the anode and collector/receptacle [22, 56]. The given formulas are

suitable for calculating the parameters of amplifier circuits at strengthening of low signals. However, it is possible to utilize them for calculating the diagrams during amplification of large signals, if we consider conductivity changes, which occur in the real amplifier instruments with a change of the input signal in ShDD.

For the formation of FAKh the conductivities must change according to the completely specific law. Utilizing expressions (2-5) and (2-8) for the voltage amplification factors of the most widely used diagrams (with the common cathode and emitter), let us find the required law of conductivity change in the load circuit $Y = f(U_{ax}, S)$ in the dynamic s-band the account a change in slope/transconductance S and, on the contrary, the law of a change in slope/transconductance $S = \varphi(U_{ax}, Y)$ with the account a change in conductivity Y .

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For the vacuum-tube amplifiers the effect of stray capacitances on the medium frequencies can be disregarded/neglected and in formula (2-5) accepted $Y_{a.s.} = 0$. Then expressions (2-5) and (2-8) can be written in general form

$$k = \frac{\partial U_{ax}}{\partial U_{ax}} = - \frac{Y_m(U_{ax})}{Y(U_{ax})}, \quad (2-11)$$

where $Y_{a.s.} = S$; $Y = Y_s + Y_{nary}$ - for the vacuum-tube amplifiers;

$Y = Y_1 + Y_{\text{nar}}$ for the transistor amplifiers.

Amplitude characteristic of FU is described by the expression

$$\dot{U}_{\text{out}} = -\dot{U}_{\text{in}} \frac{Y_{21}}{Y}. \quad (2-12)$$

Differentiating equation (2-12), we obtain

$$b = \frac{\partial U_{\text{out}}}{\partial U_{\text{in}}} = -U_{\text{out}} \cdot \frac{\partial K}{\partial U_{\text{in}}} + K, \quad (2-13)$$

where

$$\frac{\partial K}{\partial U_{\text{in}}} = \frac{Y Y_{21} + Y_{21} Y'}{Y^2}.$$

Utilizing expressions for the differential factor of amplifications b , given in table 1, and equation (2-13), we find required dependences $Y = f(U_{\text{in}}, Y_{21})$ and $Y_{21} = \varphi(U_{\text{in}}, Y)$ for the realization of different forms of FAKh.

Since the course and the general solution during finding of the required dependences are identical in the implementation of any form of FAKh, let us examine the course of solution of assigned mission based on the example to realization of LAKh.

Substituting the value differential gear ratio/transmission factor for the amplifier with LAKh to equation (2.13) and disregarding a change in the phase of output stress/voltage on 180° , i.e., omitting minus sign in expression (2-12), we obtain the following differential equation

$$Y'_{21} + Y'_{21} Y Y^{-1} + Y_{21} U_{\text{in}}^{-1} - M Y U_{\text{in}}^{-1} = 0. \quad (2-14)$$

We find dependence $Y = f(U_{\text{ex}}, Y_{\text{in}})$. For this equation (2-14) we copy/rewrite in the following form:

$$Y' + YP(U_{\text{ex}}) + Y^2Q(U_{\text{ex}}) = 0, \quad (2-15)$$

where

$$P(U_{\text{ex}}) = \frac{Y_{\text{in}}}{Y_{\text{in}}^2} - \frac{1}{U_{\text{ex}}}; \quad Q(U_{\text{ex}}) = \frac{M}{Y_{\text{in}} U_{\text{ex}}^2}.$$

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Equation (2-15) is the Bernoulli equation, which is reduced to the linear equation during the introduction to new variable $\xi = Y^{-1}$. The general solution of equation (2-15) is written/recorded in the following form:

$$Y = \left\{ \exp \left(\int_{U_{\text{ex. n}}}^{U_{\text{ex}}} P(U_{\text{ex}}) dU_{\text{ex}} \right) \left[\int_{U_{\text{ex. n}}}^{U_{\text{ex}}} Q(U_{\text{ex}}) \times \right. \right. \\ \left. \left. \times \exp \left(\int_{U_{\text{ex. n}}}^{U_{\text{ex}}} P(U_{\text{ex}}) dU_{\text{ex}} \right) dU_{\text{ex}} + C \right] \right\}^{-1}. \quad (2-16)$$

On the basis of initial conditions $Y = Y_{\text{n}}, Y_{\text{in}} = Y_{\text{in}}$ when $U_{\text{ex}} = U_{\text{ex. n}}$, we find integration constant $C = Y_{\text{n}}^{-1}$. After manufacturing integration in the limits from $U_{\text{ex. n}}$ to U_{ex} , we obtain the resultant expression

$$Y = Y_{\text{n}} \cdot \frac{Y_{\text{in}}}{Y_{\text{in}}^2} \cdot \frac{s}{s}, \quad (2-17)$$

where Y_{n} and Y_{in} - initial values of conductivities Y and Y_{in} with the work of amplifier in the linear conditions.

Solving expression (2-17) relative to Y_{in} , we find

$$Y_{\text{in}} = Y_{\text{in}} \cdot \frac{Y}{Y_{\text{n}}} \cdot \frac{s}{s}. \quad (2-18)$$

and

Laws (2-17)^(2-18) general/common/total and are valid for the realization as functional, so to quasi-linear of the modes/conditions of the work of amplifier with any FAKh with the substitution of the corresponding expressions for the standardized/normalized stresses/voltages x and z .

Conductivity Y_{11} is slope/transconductance S of the passage characteristic of amplifier instrument and changes under the effect of voltage input, while conductivity Y - under the effect of output stress/voltage.

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Therefore it is expedient to examine the dependences

$$Y(z) = Y_{11} \Phi(S) \varphi(z); \quad (2-19)$$

$$Y_{11}(x) = S(x) = S_{11} \Phi(Y) \psi(x), \quad (2-20)$$

where $\Phi(S) = \frac{S}{S_{11}}$ - function, which characterizes a change in the slope/transconductance in the dynamic range; $\Phi(Y) = \frac{Y}{Y_{11}}$ - function, which characterizes a change in the load admittance in the dynamic range; $\varphi(z) = x(z)/z$ - ratio x/z , expressed in the standardized/normalized units z ; $\psi(x) = z(x)/x$ - ratio z/x , expressed in the standardized/normalized units x . The functions $\varphi(z)$ and $\psi(x)$, which ensure the realization of different types of FAKh, are given in Table 5.

With the constants to slope/transconductance $S = S_n = \text{const}$ and load $Y = Y_n = \text{const}$ is satisfied the condition $F(S)=1$ and $F(Y)=1$.

Dependences of form (2-19) can be utilized for calculating the amplifiers, in which FAKh is realized due to the nonlinear properties of load taking into account nonlinearity of UP.

Dependences of form (2-20) can be utilized for calculating the amplifiers, in which FAKh realize due to the nonlinear properties of amplifier or due to the intentional change in conductivity Y_{11} (slope/transconductance S) taking into account the nonlinearity of load (or the nonlinearity of input admittance of the following cascade/stage).

Table 5.

(2) Тип ФАХ	(1) Функция			
	$\varphi(z)$		$\psi(z)$	
	(3) Функциональ- ная	(4) Квазилинейная	(3) Функциональ- ная	(4) Квазилинейная
(5) Логарифми- ческий	$\exp \frac{z-1}{a}$	$\left(\frac{z-1}{a} - \ln a + 1\right) d$	$\frac{a \ln z + 1}{z}$	$a \left(\ln d + \frac{z}{d} - 1\right) + 1$
(6) Экспонен- циальная	$\frac{a \ln z + 1}{z}$	(7) Нет	$\exp \frac{z-1}{a}$	(7) Нет
(8) Степенная	$\beta = \frac{1}{n}$	$\left(\frac{z-1+n}{n}\right)^n \left[z - (n-1) K_n^{\frac{1}{n-1}} + \right.$ $\left. + (n-1) K_n^{\frac{1}{n-1}} \right]$	$\frac{n \sqrt[n]{z-n+1}}{z}$	$\frac{(n-1) K_n^{\frac{1}{n-1}} + \frac{z}{K_n} - n + 1}{z}$
	$\beta = m$	(7) Нет	$\frac{z^m + m - 1}{mz}$	(7) Нет

Key: (1). Function. (2). Type of FAKh. (3). Functional. (4). Quasi-linear. (5). Logarithmic. (6). Exponential. (7). No. (8). Exponential.

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§2. Characteristics and the parameters of amplifier instruments and nonlinear elements/cells.

The realization of the prescribed/assigned form of FAKh depends on the form of the passage characteristic of amplifier instrument $i_{\text{max}} = j(\dot{U}_{\text{ax}})$ and current-voltage characteristic of nonlinear

element/cell $i_{\text{non}} = f(U_{\text{non}})$. For the realization of FAKh it is necessary to know the basic parameters of UP and the nonlinear elements/cells: the slope/transconductance of the passage characteristic S of amplifier instrument; conductivity g_{non} or resistance R_{non} of nonlinear element/cell.

During the development of diagrams of FU it is sometimes more convenient to operate with resistance, since in the catalogs of parts is given resistance, but not conductivities.

The enumerated parameters of real instruments change with a change in the signal and have the different value under the influence of different signal aspects. Therefore should be distinguished three forms of the parameters indicated: differential slope/transconductance, conductivity and resistance, which can be used for the calculation of linear amplifiers, for the amplifier instrument

$$S = \frac{di_{\text{non}}}{dU_{\text{non}}}; \quad (2-21)$$

for the nonlinear element/cell

$$g_{\text{non}} = \frac{1}{R_{\text{non}}} = \frac{di_{\text{non}}}{dU_{\text{non}}}; \quad (2-22)$$

average/mean slope/transconductance S_{cp} , conductivity $g_{\text{non, cp}}$ and resistance $R_{\text{non, cp}}$ during strengthening of pulse signal, determined by the method of secant (Fig. 27)

$$S_{cp}^{(\pm)} = \frac{I_{\text{вых}}^{\pm}}{U_{\text{вх}}^{\pm}}; \quad (2-23)$$

$$g_{\text{нел. ср}}^{\pm} = \frac{I_{\text{нел}}^{\pm}}{U_{\text{нел}}^{\pm}}, \quad (2-24)$$

where $I_{\text{вых}}^{\pm}$ - value of output current respectively for positive and negative pulse signal U^{\pm} ; $I_{\text{нел}}^{\pm}$ - the current, flowing through the nonlinear element/cell, caused by the effect of positive or negative pulse signal; average/mean slope/transconductance, conductivity and resistance in the amplitude of the fundamental harmonic of the output current of amplifier instrument or current, flowing through the nonlinear element/cell, in the case of strengthening the harmonic oscillations

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The values of parameters S , $g_{\text{нел}}$, $R_{\text{нел}}$ depend on level and form of the amplified signal and on the form of passage characteristic of UP and volt-ampere characteristic of nonlinear element/cell.

Passage characteristic of UP can be represented as the function of output (anodic, collector) current from control voltage $i = f(e_g)$, which can be referred either to the grid (base), or to the anode (to the collector/receptacle).

$$S_{cp1} = \frac{I_{m1}}{U_m}; \quad (2-25)$$

$$R_{\text{нел1}} = \frac{1}{g_{\text{нел1}}} = \frac{U_m}{I_{m1}}. \quad (2-26)$$

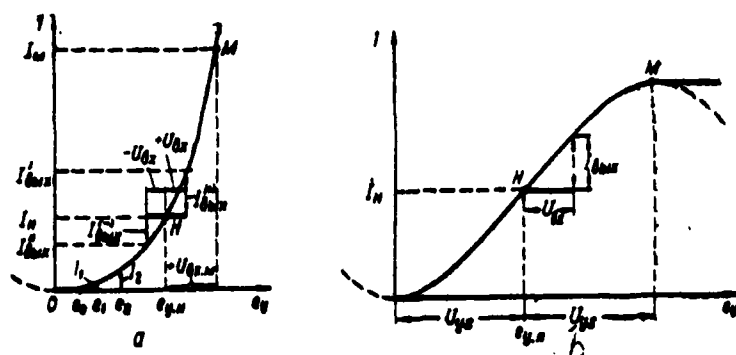


Fig. 27. Approximation of the passage characteristic of amplifier instrument by the polynomial: a) the second degree; b) the third degree.

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Table 6.

(1) Аппроксимация вольт-ампер- ной характеристики $i = f(U)$	(2) Дифференциальная крутизна $S_{\text{диф}} = \frac{\partial i}{\partial U} = f'(U)$	(3) Обратная ампер-вольтная характеристика $U = f(i)$	(4) Дифференциальное сопротивление $R_{\text{диф}} = \frac{1}{S_{\text{диф}}} = \frac{\partial U}{\partial i}$
(5) Полиномиальная $i = a_0 + a_1 U + a_2 U^2 + a_3 U^3 + \dots$ $\dots = \sum_{k=0}^n a_k U^k$	$a_1 + 2a_2 U + 3a_3 U^2 + \dots$	$U = \frac{i}{a_0} - \frac{a_1}{a_0^2} i^2 + \frac{2a_2 - a_0 a_3}{a_0^3} i^3 \times$ $\times i^2 + \dots$	$\frac{1}{a_0} - \frac{2a_1}{a_0^2} i + 3 \frac{2a_2 - a_0 a_3}{a_0^3} i^2 + \dots$
(6) Экспоненциальная $i = i_0 (e^{\gamma U} - 1)$ (полупроводниковый диод, транзистор)	$\gamma i_0 e^{\gamma U} = \gamma i$	$U = \frac{1}{\gamma} \ln \left(\frac{i}{i_0} + 1 \right)$	$\frac{1}{\gamma i_0} e^{\gamma U}$
(7) $i = i_0 e^{\gamma U}$ (вакуумный диод, лампа)	,	$U = \frac{1}{\gamma} \ln \frac{i}{i_0}$,
(8) Степенная $i = A U^n$	$n A U^{n-1}$	$U = \left(\frac{i}{A} \right)^{\frac{1}{n}}$	$\frac{U^{1-n}}{n A}$
(9) Гиперболический тангенс $i = I_0 (1 + \tanh qU)$	$\frac{I_0}{\cosh^2 qU}$	$U = \frac{1}{q} \operatorname{Arth} \left(\frac{i}{I_0} - 1 \right)$	$\frac{I_0^2}{qi (2I_0 - i)}$
(10) Кусочно-параболическая $i = \begin{cases} 0 & \text{при } U < U_0 \\ \frac{1}{b} (U - U_0)^n & \text{при } U > U_0 \end{cases}$	$\frac{0}{b(U - U_0)^{n-1}}$	$U = \sqrt[n]{\frac{i}{b}} + U_0$	(11) $\frac{\infty}{b(U - U_0)^{n-1}}$ при $U < U_0$, $\frac{1}{b(U - U_0)^{n-1}}$ при $U > U_0$.

Key: (1).

Approximation of volt-ampere characteristic. (2). Differential slope/transconductance. (3). Reverse/inverse ampere-volt characteristic. (4). Differential resistance. (5). Poly-nominal. (6). Exponential (semiconductor diode, transistor). (7). (vacuum-tube diode, tube). (8). Exponential. (9). Hyperbolic tangent. (10). piecewise-parabolic. (11). with.

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Is greatly extended the determination of the current through the control voltage of grid (base), which is equal to:

for the vacuum triodes

$$e_y = U_0 + DU_{a1} \quad (2-27)$$

for the transistors

$$e_y = U_{0.0} + DU_{a1} \quad (2-28)$$

for the tetrodes

$$e_y = U_0 + D_1 U_{a2} + D_1 D_2 U_{a1} \quad (2-29)$$

for the pentodes

$$e_y = U_0 + D_1 U_{a2} + D_1 D_2 U_{a2} + D_1 D_2 D_3 U_{a1} \quad (2-30)$$

where D_1 , D_2 , D_3 - permeability respectively on the controlling/guiding (the first), shielding (the second) and shielding (the third) grids; D - permeability on the control electrode of triode. Formula for determining of D for the transistor is given in Appendix 2.

Products $D_1 D_2$ and $D_1 D_2 D_3$ are so low that for the tetrodes and the pentodes it is possible to count

$$e_y = U_0 + D_1 U_{a2} \approx U_0 \quad (2-31)$$

Dependence $i = f(e_y)$ for real UP is nonlinear. The most widely used functions, by using which it is possible to sufficiently accurately

approximate the volt-ampere characteristics of nonlinear elements/cells, are given in Table 6. The same table gives expressions for the differential values of the slope/transconductance (conductivity) of characteristic and internal resistance of nonlinear element/cell.

The possible approximation methods of the characteristics of nonlinear elements/cells are in sufficient detail examined in the work of the Soviet [4, 5, 23, 31, 32] and foreign [74, 92, 107] authors. Should be especially noted generalizing work [5]. However, in these works, besides [23], is not examined average/mean steepness of the characteristic and average/mean conductivity of nonlinear element/cell. Therefore it is necessary at least to briefly examine the most widely used methods of approximation, to estimate these methods from the point of view of convenience in the analytical determination of average/mean slope/transconductance and to find the analytical expressions, which are determining the dependence of the average/mean slope/transconductance of passage characteristic of UP on the value of pulse and harmonic signal,.

POLYNOMINAL APPROXIMATION.

The polynomial of the n degree, which approximates characteristic, in the convoluted form can be written

$$i = \sum_{k=0}^n a_k U^k, \quad (2-32)$$

where n - degree of polynomial; a_k - coefficient, which has dimensionality $[a \cdot e^{-k}]$; k - number of component/term/addend.

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It is possible to approximate characteristic with any given precision/accuracy by the polynomial of the n degree. The precision/accuracy of approximation grows/rises with an increase in the members of polynomial. Into expression (2-32) it enters $n+1$ coefficients a_k , whose value can be selected so that the approximate and precise values of currents would coincide at $n+1$ points, called interpolation points. If interpolation units e_0, e_1, \dots, e_n are selected arbitrarily, dependence $i = f(e_y)$ can be written in the form of the interpolation polynomial of Lagrange [34]

$$i = \sum i_k \frac{(e - e_0) \dots (e - e_{k-1})(e - e_{k+1}) \dots (e - e_n)}{(e_n - e_0) \dots (e_n - e_{k-1})(e_n - e_{k+1}) \dots (e_n - e_n)}, \quad (2-33)$$

where i_k - value of current in the k interpolation unit when $e = e_k$. The zero unit e_0 can be selected, also, when $e_y = 0$.

With even pitch of interpolation

$$\Delta e_k = e_k - e_{k-1} = \Delta e_{k+1} = e_{k+1} - e_k = \text{const}$$

interpolation polynomial can be written in the form of the Newton polynomial [24]

$$i = i(e_0) + \sum_{k=1}^n \frac{\Delta^k i(e_0)}{k! \Delta e^k} (e - e_0)(e - e_0 - \Delta e) \dots [(e - e_0 - k - 1) \Delta e], \quad (2-34)$$

where $\Delta^k i(e_0)$ - k difference, which linearly is expressed as the values of current in the interpolation points i_k with the aid of binomial coefficients C_k^m

$$\Delta^k i(e_0) = \sum_{m=0}^k (-1)^{k-m} C_k^m i_m, \quad C_k^m = \frac{k!}{m!(k-m)!}.$$

As can be seen from expressions (2-33) and (2-34), the polynomials of Lagrange and Newton are not polynomials according to degrees e , therefore the dependence of average/mean slope/transconductance S_{cp} on control voltage $S_{cp} = f(e_y)$ cannot be written analytically. Furthermore, for obtaining the high precision/accuracy it is necessary to take a sufficiently large quantity of interpolation units, which leads to the polynomial of high degree. However, with a number of terms more than three-four mathematical analysis of FU sharply become complicated.

The degree of polynomial can be lowered during the determination of the coefficients of the polynomial in the method of least squares, which leads to the solution of the system of canonical equations.

In work [31] it is proposed instead of the direct solution of canonical system of equations to solve the system of orthogonal polynomials, which possesses the positive property, that with the expansion in terms of them of any function the best approximation in the sense of the least squares is obtained. However, in this case approximated that is not expressed explicitly analytical depending on control voltage e_y , therefore dependence $S_{op} = f(e_y)$ cannot be written analytically also.

It is easy to find this dependence, if for determining the terms of polynomial to use the expansion of function $i_{max}(e_{y, n} + \Delta e_y)$ in the Taylor series, according to degrees Δe_y

$$i(e_{y, n} + \Delta e_y) = i(e_{y, n}) + \left(\frac{di}{de_y}\right)_{e_{y, n}} \Delta e_y + \frac{1}{2!} \left(\frac{d^2i}{de_y^2}\right)_{e_{y, n}} \Delta e_y^2 + \dots + \frac{1}{3!} \left(\frac{d^3i}{de_y^3}\right)_{e_{y, n}} \Delta e_y^3 + \dots = i_n + \alpha U_{in} + \beta U_{in}^2 + \gamma U_{in}^3 + \dots, \quad (2-35)$$

where $U_{in} = \Delta e_y$ - the voltage input of signal; $\alpha = S$; $\beta = (1/2!)S'$; $\gamma = (1/3!)S''$; ..., S, S', S'' - slope/transconductance and its derived characteristics $i(e_y)$ at point $e_y = e_{y, n}$.

A relative value of separate members and their number in series/row (2-35) with the prescribed/assigned precision/accuracy of approximation are determined by the position of operating point on the characteristic of UP and by the maximum value of voltage input.

If operating point on characteristic is arranged/located asymmetric, when to the positive increases Δe_y correspond increases in the output current greater, than negative ones, in the series/row it suffices to consider three members. In this case

$$i_{\text{max}} = i_n + aU_{\text{ex}} + \beta U_{\text{ex}}^2. \quad (2-36)$$

In equation (2-36) the coefficient β is unknown, which can be determined as follows. For maximum positive input signal $+U_{\text{ex. M}}$ the steepness of the characteristic S_M at point M (Fig. 27a) is determined graphically. Then the is performed identity

$$\left(\frac{di}{dU_{\text{ex}}}\right)_{U_{\text{ex}} = U_{\text{ex. M}}} = S_M = a + 2\beta U_{\text{ex. M}} = S_n + 2\beta U_{\text{ex. M}}$$

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Whence

$$\beta = S_n \frac{(\eta - 1)}{2U_{\text{ex. M}}}, \quad (2-37)$$

and equation (2-36) can be written

$$i_{\text{max}} = i_n + S_n U_{\text{ex}} + S_n \frac{(\eta - 1)}{2U_{\text{ex. M}}} U_{\text{ex}}^2, \quad (2-38)$$

where $\eta = \frac{S_M}{S_n}$; S_n - slope/transconductance at point H when $e_y = 0$.

Differential slope/transconductance

$$S = \frac{\partial i}{\partial U_{\text{ex}}} = S_n [1 \pm (\eta - 1)q], \quad (2-39)$$

where $q = \frac{U_{\text{ex}}}{U_{\text{ex. M}}}$.

Pulse positive signal.

$$S_{\text{op}}^{(+)} = \frac{i_{\text{max}}^{(+)}}{U_{\text{ex}}^{(+)}} = a + \beta U_{\text{ex}} = S_n \left(1 + \frac{\eta - 1}{2} q\right) = S_n \Phi(S^+), \quad (2-40)$$

where $I_{\text{max}}^{(+)} = I'_{\text{max}} - I_{\text{max. n}}$ - increase in the output current, caused by pulse positive signal; $\Phi(S^+) = \frac{S_{\text{cp}}^{(+)}}{S_{\text{n}}} = \left(1 + \frac{\eta-1}{2}q\right)$ - function, which characterizes a change in the steepness of the characteristic in the dynamic range.

Pulse negative signal. Analogously we obtain

$$S_{\text{cp}} = \frac{I_{\text{max}}^{(-)}}{U_{\text{ax}}^{(-)}} = S_{\text{n}} \Phi(S^-), \quad (2-41)$$

where

$$\Phi(S^-) = 1 - \frac{\eta-1}{2}q.$$

Harmonic signal $U_{\text{ax}} = U_{\text{m}} \cos \omega t$. Substituting in equation (2-38), we obtain

$$i_{\text{max}} = I_{\text{n}} + I_{\text{m}_1} \cos \omega t + I_{\text{m}_2} \cos 2\omega t, \quad (2-42)$$

$$I_{\text{n}} = i_{\text{n}} + \frac{S_{\text{n}}}{4} \cdot \frac{\eta-1}{U_{\text{ax. n}}}; \quad (2-43)$$

$$I_{\text{m}_1} = S_{\text{n}} U_{\text{m}}; \quad (2-44)$$

$$I_{\text{m}_2} = \frac{S_{\text{n}}}{4} \cdot \frac{\eta-1}{U_{\text{ax. n}}} - \quad (2-45)$$

respectively constant component and the amplitude of the first and second harmonics of output current.

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Average/mean slope/transconductance in amplitude I_{m}

$$S_{\text{cp1}} = \frac{I_{\text{m}_1}}{U_{\text{m}}} = S_{\text{n}} = \text{const} \quad (2-46)$$

is constant value and does not depend on input signal level.

Standardized/normalized dependences $\phi(S^{\pm}) = f(q)$ for the positive, the negative and the harmonic of input signals at the different values η are given in Fig. 28a. In the particular case with $\eta=4$

$$U_{2k, n} = U_{7, n}.$$

If operating point is established/installed on the middle of symmetrical characteristic (Fig. 27b), then even degrees in the decomposition/expansion drop out and characteristic can be approximated with the aid of the polynomial, which contains only the odd degrees

$$i_{2k+1} = i_n + aU_{2k} + \gamma U_{2k}^3 + \dots \quad (2-47)$$

For guaranteeing the mathematical analysis it is possible to be restricted to the polynomial of the third power

$$i_{2k+1} = i_n + aU_{2k} + \gamma U_{2k}^3. \quad (2-48)$$

Characteristic corresponding to this approximation is shown by prime in Fig. 27b.

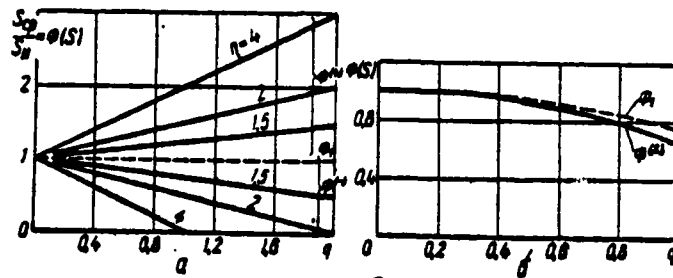


Fig. 28. Graph/diagrams of dependences $F(S)=f(q)$ with the approximation of characteristic of UP by the polynomial: a) the second degree; b) the third degree.

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In expression (2-48) the coefficient γ is unknown, which can be determined as follows. According to the objective parameter the value of saturation voltage U_{ys} is determined at which the steepness of the characteristic is equal to zero (Fig. 27b). In this case the identity

$$\left(\frac{\partial i}{\partial U_{\gamma}}\right)_{U_{\gamma}=U_{ys}} = \alpha + 3\gamma U_{ys} = 0,$$

is performed whence

$$\gamma = -\frac{\alpha}{3U_{ys}^2} = -\frac{S_n}{3U_{ys}^2}. \quad (2-49)$$

Substituting the value γ into formula (2-47), we obtain

$$i_{max} = i_n + S_n U_{nx} \left(1 - \frac{q^2}{3}\right), \quad (2-50)$$

where

$$q = \frac{U_{nx}}{U_{ys}} = \frac{U_{nx}}{U_{y2}}, \quad .$$

since $U_{\text{ex. n}} = U_{\text{ys.}}$

Differential slope/transconductance

$$S = \frac{\partial i_{\text{max}}}{\partial U} = S_n (1 - q^2) = S_n \Phi(S). \quad (2-51)$$

Since the characteristic is symmetrical relative to working point, average/mean slope/transconductance changes equally for the positive and negative pulse of the signals

$$S_{\text{op}}^{(\pm)} = S_n \left(1 - \frac{q^2}{3}\right) = S_n \Phi(S \pm). \quad (2-52)$$

With the harmonic signal expression (2-50) takes the form

$$i_{\text{max}} = I_n + I_{m1} \cos \omega t + I_{m3} \cos 3\omega t, \quad (2-53)$$

where respectively constant component and amplitudes of the first and third harmonics of output current are determined by the expressions:

$$I_n = i_n; \quad (2-54)$$

$$I_{m1} = S_n U_m \left(1 - \frac{q^2}{4}\right); \quad (2-55)$$

$$I_{m3} = U_m S_n \frac{q}{3 \cdot 4}. \quad (2-56)$$

Average/mean slope/transconductance in amplitude I_{m1}

$$S_{\text{op}} = \frac{I_{m1}}{U_m} = S_n \left(1 - \frac{q^2}{4}\right) = S_n \Phi(S). \quad (2-57)$$

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Plotted functions Φ^* for the present instance are depicted in Fig. 28b. The approximation of objective parameter is permitted by the polynomial of third power (Fig. 27b) with $q \leq 1$. From the graph Fig. 28b shows that with $q \leq 0.4$ change $S_{op}^{(\pm)}$ does not exceed 5%, and $S_{op} - 4\%$, with $q \leq 0.4$ amplifier instrument can be considered virtually linear.

EXPONENTIAL APPROXIMATION.

By exponential function of the form

$$i = i_0 \exp \gamma e_{\gamma}, \quad (2-58)$$

where i_0 - current when $e_{\gamma} = 0$; γ - coefficient, whose value varies in limits of $5-10 \text{ V}^{-1}$, sufficiently accurately it is possible to approximate the characteristics of vacuum devices (diodes, multielectrode tubes) with the negative voltages on control electrode.

Let us pause at the mathematical description of the objective parameters of semiconductor devices (transistors and diodes) and

their approximation.

Transistor is inertial UP, since the charge carriers in the region of base are moved with finite time. Disregarding the inertial properties for planar type germanium transistors, for which it is possible to consider that the admixtures/impurities in the base are distributed evenly, and the geometry is close to parallel, volt-ampere characteristics when $U_{be} > 0.5 \text{ V}$ in according to works [42, 66] have the following mathematical recording:

$$I_e = A_e [e^{\lambda U_{be} / (kT)} - 1]; \quad (2-59)$$

$$I_u = A_u [e^{\lambda U_{be} / (kT)} - 1]; \quad (2-60)$$

$$I_o = A_o [e^{\lambda U_{be} / (kT)} - 1], \quad (2-61)$$

where A_e, A_u, A_o - coefficients, which have the dimensionality of current and being the functions of the properties of the material, from which the transistor, and its geometries is prepared; $\lambda = \frac{q}{kT}$ - coefficient; q - electron charge; k - Boltzmann constant; T - absolute temperature; r_b - distributed resistance of base. At a room temperature $t = +20^\circ\text{C}$ $\lambda = 0.025 \text{ V}^{-1}$.

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The input characteristic of the transistor, connected on the common-base circuit, is described by expression (2-59), by expression (2-61) - the input characteristic of the transistor, connected on the diagram with O3, and (2-60) - passage characteristic of the

transistor, connected both on the diagram with O3 and with O5.

Since for the transistor the equality

$$I_0 = I_0 - I_K = I_0 (1 - \alpha_0), \quad (2-62)$$

is implemented where $\alpha_0 = \frac{\Delta I_K}{\Delta I_0}$ and $B = \frac{\Delta I_K}{\Delta I_0} = \frac{\alpha_0}{1 - \alpha_0}$ - static current amplification factors with respect to common-base circuits and emitter with the short circuit at the output, it is possible to register

$$A_0 = A_0 (1 - \alpha_0); \quad (2-63)$$

$$A_K = A_0 B. \quad (2-64)$$

The values of coefficient A_0 and resistor/resistance r_0 for the junction transistors of different types vary in the sufficiently wide limits. For the high-frequency transistors of types P401-P403, P411, P415, P416, P418Zh the value of coefficient A_0 is within limits of $(0.5-2) \cdot 10^{-4}$ and resistance r_0 can have a value from 30 to 300-500 ohms.

The volt-ampere characteristic of semiconductor diode is hack-written by the dependence, analogous (2-59)-(2-61)

$$i_{\text{нес}} = A [e^{\lambda(U_{\text{нес}} - i_{\text{нес}} r)} - 1], \quad (2-65)$$

where r - volumetric resistor/resistance of semiconductor. Expression (2-59), (2-60), (2-61) and (2-65) transcendental ones, from cannot be used directly for calculating the average/mean values of parameters S , $g_{\text{нес}}$ and $R_{\text{нес}}$. According to the given expressions it is possible to determine only the differential parameters. Using expression (2.65),

let us consider two cases:

1. Current i_{ges} is small and is fulfilled inequality

Then $i_{\text{ges}} r \ll U_{\text{ges}}$.

$$i_{\text{ges}} = A \exp(\lambda U_{\text{ges}} - 1).$$

Differential slope/transconductance

$$S = \frac{\partial i}{\partial U} = \lambda A e^{\lambda U_{\text{ges}}} \approx \lambda i_{\text{ges}}. \quad (2-66)$$

2. Current i_{ges} is great it is implemented inequality

$\exp[\lambda(U_{\text{ges}} - i_{\text{ges}} r)] \gg 1$. Then

$$i_{\text{ges}} = A \exp[\lambda(U_{\text{ges}} - i_{\text{ges}} r)]. \quad (2-67)$$

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We take the logarithm of expression (2-67) and solve relatively

U_{ges}

$$U_{\text{ges}} = i_{\text{ges}} r + \frac{1}{\lambda} (\ln i_{\text{ges}} - \ln A).$$

We find the partial derivative

$$r_{\text{ges}} = \frac{\partial U_{\text{ges}}}{\partial i_{\text{ges}}} = r + \frac{1}{\lambda i_{\text{ges}}}. \quad (2-68)$$

Hence we obtain the value of differential slope/transconductance (or input admittance)

$$S = g_{\text{ges}} = \frac{1}{r_{\text{ges}}} = \frac{\lambda i_{\text{ges}}}{1 + \lambda i_{\text{ges}} r}. \quad (2-69)$$

If we into expression (2-60) substitute $I_0 = \frac{I_n}{B}$ and to designate

$\frac{r_0}{B} = a$ for the passage characteristic of the transistor, connected on the diagram with O3, we will obtain

$$S = g_{21} = \frac{1}{r_{21}} = \frac{M_K}{1 + \lambda M_K}. \quad (2-70)$$

For facilitating the calculation of the average parameters and analysis of nonlinear diagram each of expressions (2-59), (2-60), (2-61) and (2-65) is expedient to replace with the approximating function of the form

$$i = i_0 e^{\gamma U} \quad (2-71)$$

or form (2-58).

For transistors $U = U_{be}$, for diodes $U = U_{an}$. The value of coefficients of i_0 and γ can be determined in the experimentally taken/removed characteristic. For this it is necessary to assign two values of current i_1 and i_2 also, according to the experimental characteristic $i = f(U)$ to determine the values of U_1 corresponding to them and U_2 . According to expression (2-58)

$$i_1 = i_0 e^{\gamma U_1}; \quad i_2 = i_0 e^{\gamma U_2}.$$

Solving equations relatively γ and i_0 , we obtain

$$\gamma = \frac{1}{U_2 - U_1} \ln \frac{i_2}{i_1}, \quad i_0 = A = \frac{i_1}{\gamma U_1}.$$

The value of differential slope/transconductance is given in Table 6. Initial value of slope/transconductance when $U_{bx} = 0$

$$S_0 = \lim_{U_{bx} \rightarrow 0} \frac{\partial i(\pm)}{\partial U_{bx}} = \gamma i_0 e^{\gamma U_0} = \gamma i_0. \quad (2-72)$$

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Then average/mean slope/transconductance for the positive and negative pulse signals with given one $e_{\gamma, n}$ according to Fig. 27

$$S_{op}^{(\pm)} = \frac{S_n}{\gamma U_{nx}} (\pm e^{\gamma U_{nx}} \mp 1) = S_n \Phi(S_{\pm}). \quad (2-73)$$

As can be seen from expression (2-73) average/mean slope/transconductance with an increase in the positive signal increases and, on the contrary, it is reduced for the negative signal.

With harmonic signal $U_{nx} = U_m \cos \omega t$

$$i_{nax} = i_0 [\exp \gamma (e_{\gamma, n} + U_m \cos \omega t) - 1] = i_0 [e^{\gamma e_{\gamma, n}} e^{\gamma U_m \cos \omega t} - 1].$$

Factor with $e^{\gamma U_m \cos \omega t}$ is the function of form $e^{x \cos \omega t}$, which is expanded into series according to the functions of Bessel of the imaginary argument

$$e^{x \cos \omega t} = J_0(jx) + 2 \sum_{n=1}^{\infty} J_n(jx) \cos n \omega t. \quad (2-74)$$

Then

$$i_{nax} = i_0 e^{\gamma e_{\gamma, n}} [J_0(jx) + 2 \sum_{n=1}^{\infty} J_n(jx) \cos n \omega t - 1].$$

After developing this series/row, we obtain

$$i_{nax} = I_0 + I_{m1} \cos \omega t + I_{m2} \cos 2\omega t + \dots, \quad (2-75)$$

where dc current component

$$I_0 = i_0 [e^{\gamma e_{\gamma, n}} J_0(jx) - 1]; \quad (2-76)$$

the amplitude of the fundamental harmonic of the current

$$I_{m1} = 2j^{-1} i_0 e^{\gamma e_{\gamma, n}} J_1(jx); \quad (2-77)$$

$j^{-1}J_1(jx)$ - the modified function of first-order Bessel; $x = \gamma U_m$.

It is known that

$$J_n(x) = j^{-1}J(jx) = \frac{\left(\frac{x}{2}\right)^n}{\Gamma(n+1)} + \frac{\left(\frac{x}{2}\right)^{n-2}}{\Gamma(n+2)} + \frac{\left(\frac{x}{2}\right)^{n-4}}{2!\Gamma(n+3)} + \dots + \frac{\left(\frac{x}{2}\right)^{2n}}{\left(\frac{n}{2}\right)!\Gamma(2n)}; \quad (2-78)$$

$$j^{-1}J_1(jx) = \frac{x}{2} + \frac{\left(\frac{x}{2}\right)^3}{2} + \frac{\left(\frac{x}{2}\right)^5}{2!3!} + \dots \quad (2-79)$$

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After substituting expression (2-79) in (2-78) taking into account, that $x = \gamma U_m$, we obtain

$$I_{m1} = I_0 e^{\gamma U_m} = \left[\gamma U_m + \frac{\gamma^3 U_m^3}{2^3} + \frac{\gamma^5 U_m^5}{3 \cdot 2^3} + \dots \right]. \quad (2-80)$$

Average/mean slope/transconductance

$$S_{op1} = \frac{I_{m1}}{U_m} = I_0 e^{\gamma U_m} = \left[\gamma + \frac{\gamma^3 U_m^2}{2^3} + \frac{\gamma^5 U_m^4}{3 \cdot 2^3} + \dots \right]. \quad (2-81)$$

Taking into account (2-66), we obtain

$$S_{op1} = S_n \left(1 + \frac{\gamma^2 U_m^2}{2^3} + \frac{\gamma^4 U_m^4}{3 \cdot 2^3} + \dots \right). \quad (2-82)$$

For the contemporary transistors the value of coefficient γ can oscillate from 5 to 40 (ideal transistor, which has $r_0 = 0$). Therefore the calculation of slope/transconductance according to formula (2-82)

is sufficiently precise with the limited number of terms only for small amplitudes.

With large amplitudes ($x \gg n$) instead of series/row (2-78) it is possible to use the asymptotic expression

$$J_n(x) = \frac{e^x}{\sqrt{2\pi x}} \sum_{k=0}^{\infty} \frac{\left(\frac{1}{2} - n\right)_k \left(\frac{1}{2} + n\right)_k}{k! (2x)^k}, \quad (2-83)$$

where in the numerator by brackets with index k are designated the products of form $(p)_k = p(p+1) \dots (p+k-1)$, $p_0 = 1$. Since $x = \gamma U_m \gg n$, in expression (2-83) it is possible to be bounded by two-three members

$$J_n(x) \approx \frac{e^x}{\sqrt{2\pi x}} \left[1 + \frac{1-4n^2}{8x} + \frac{(1-4n^2)(9-4n^2)}{2(8x)^2} + \dots \right].$$

Then for the amplitude of the fundamental harmonic of current and average/mean slope/transconductance

$$I_m = \frac{\sqrt{2} I_0 \gamma e^{\gamma U_m}}{\sqrt{\pi \gamma U_m}} \left(1 - \frac{3}{8 \gamma U_m} - \frac{3 \cdot 5}{2 (8 \gamma U_m)^2} - \dots \right); \quad (2-84)$$

$$S_{cp1} = \frac{\sqrt{2} I_0 S_n \exp \gamma U_m}{(\gamma U_m)^{1/2} \sqrt{\pi}} \left(1 - \frac{3}{8 \gamma U_m} - \frac{3 \cdot 5}{2 (8 \gamma U_m)^2} - \dots \right). \quad (2-85)$$

When $\gamma U_m \gg 1$ in expression (2-84) all terms in the brackets in comparison with one can be disregarded/neglected. Then

$$I_{m1} \approx \frac{\sqrt{2} I_0 \exp(\gamma(e_{\gamma, n} + U_m))}{\sqrt{\pi \gamma U_m}}. \quad (2-86)$$

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Average/mean slope/transconductance

$$S_{cp1} = \frac{\sqrt{2} S_n \exp \gamma U_m}{U_m \sqrt{\pi \gamma U_m}}. \quad (2-87)$$

Amplifier instrument with the exponential characteristic can be considered linear with satisfaction of the condition

$$\gamma U_m < 0.1.$$

Consequently, with $\gamma=10-40$ transistor can be considered linear amplifier instrument with $U_m < (10+2.5) \text{ mV}$.

Approximation of passage characteristic of UP by the hyperbolic tangent $i=I_0(1+\text{tg}qU)$.

In work [36] it is shown that sufficiently well it is possible to approximate the characteristics of a considerable number of types of tubes by hyperbolic tangent (Fig. 29).

Differential slope/transconductance

$$S = \frac{\partial i}{\partial U_{bx}} = \frac{I_0}{\text{ch}^2 q (U_{c. n} \mp U_{bx})}. \quad (2-88)$$

Average/mean slope/transconductance for the negative pulse signal according to Fig. 29

$$S_{cp}^{(-)} = \frac{I_{bx}^{(-)}}{U_{bx}} = \frac{I_0 |\text{th} q U_{c. n} - \text{th} q (U_{c. n} - U_{bx})|}{U_{bx}},$$

for the positive signal

$$S_{cp}^{(+)} = \frac{I_{bx}^{(+)}}{U_{bx}} = \frac{I_0 |\text{th} q (U_{c. n} + U_{bx}) - \text{th} q U_{c. n}|}{U_{bx}}.$$

After conversions we obtain

$$S_{cp}^{(\pm)} = \frac{I_0}{U_{BX}} \cdot \frac{\operatorname{sh} q U_{BX}}{\operatorname{ch} q U_{c. n} \operatorname{ch} [q (U_{c. n} \pm U_{BX})]}. \quad (2-89)$$

Initial slope/transconductance when

$$S_n = \lim_{U_{BX} \rightarrow 0} \frac{\partial I_{BMK}^{(\pm)}}{\partial U_{BX}} = \frac{I_0}{\operatorname{ch} q U_{c. n}}. \quad (2-90)$$

Taking into account S_n when $I_{BX} \rightarrow 0$ finally we obtain

$$S_{cp}^{(\pm)} = \frac{S_n \operatorname{sh} q U_{BX} \operatorname{ch} q U_{c. n}}{U_{BX} \operatorname{ch} q (U_{c. n} \pm U_{BX})}. \quad (2-91)$$

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With the harmonic signal it is not possible to obtain single-valued mathematical recording for the average/mean slope/transconductance. Therefore it is expedient from formula (2-87) to first calculate dependence $I_{m1} = f(U_m)$, and then - dependence $S_{cp1} = \varphi(U_m)$.

The determination of average/mean slope/transconductance under the influence of harmonic signal with the polygonal approximation of characteristic is examined in work [23].

In the case of complicated characteristic and harmonic signal it is not possible to obtain single-valued mathematical recording for the average/mean slope/transconductance. Then the amplitude of current I_{m1} can be calculated graph-analytically from the method of five ordinates

$$I_{m1} = \frac{1}{3} \left[I(+U_m) - I(-U_m) + I\left(+\frac{1}{2}U_m\right) - I\left(-\frac{1}{2}U_m\right) \right] \quad (2-92)$$

for the characteristic, asymmetric relative to operating point, or

$$I_{m1} = \frac{2}{3} \left[I(U_m) + I\left(\frac{1}{2}U_m\right) \right] \quad (2-93)$$

for the characteristic, symmetrical relative to operating point.

Through $I(+U_m)$; $I(-U_m)$; $I\left(+\frac{1}{2}U_m\right)$ and $I\left(-\frac{1}{2}U_m\right)$ are designated the currents, which correspond to positive and negative amplitudes of voltage/stress and to their halves.

Average/mean slope/transconductance is calculated from formula (2-25). expressions (2-92) and (2-93) are valid with any form of the approximation of characteristic UP.

We found expressions for determining the slope/transconductance of the passage characteristic of inertia-free UP, whose amplifier properties do not depend on frequency. During the calculation of mutual conductance of low-frequency gauge it is necessary the obtained expressions for S_{op} to multiply by the frequency factor m , i.e.

$$S_{cp. \pi} = S_{cp} m, \quad (2-84)$$

where

$$m = \frac{1}{\sqrt{1 + \left(\frac{\omega}{\omega_{rp}}\right)^2}} = \frac{S_{cp}}{S_{cp. \pi. \pi}}; \quad (2-85)$$

$S_{cp. \pi. \pi}$ — the value of average/mean slope/transconductance at the low frequency; ω_{rp} — the cut-off frequency of UP, at which average/mean slope/transconductance is reduced $\sqrt{2}$ once.

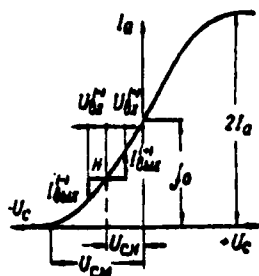


Fig. 29. Approximation of characteristic by hyperbolic tangent.

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One of the fundamental parameters for the nonlinear element/cell is average/mean conductivity or resistor/resistance, determined from formulas (2-24) and (2-26). As the nonlinear elements/cells use in essence the diodes, whose volt-ampere characteristic is approximated by exponential curve. However, supplying on the diodes the locking or triggering voltage and including them in different combinations together with the linear resistors/resistances, it is possible to obtain the most diverse forms of volt-ampere characteristics and, consequently, also different laws of a change in their conductivity and resistor/resistance.

Let us agree the combination of nonlinear element/cell (diode) together with the linear resistors/resistances to call the equivalent nonlinear element/cell (ENE), and the dependence of the value of

prompt current $i_{\text{нел.с}} = f(U_{\text{нел.с}})$, flowing through the nonlinear element/cell, then the values of the voltage of signal $U_{\text{нел.с}}$, of that applied to the nonlinear element/cell, to call dynamic volt-ampere characteristic.

Possible diagrams of ENE are given in Fig. 30a, b, c, d - ENE with the asymmetric characteristics (Fig. 31a for Fig. 30a and b); e, f - ENE with the symmetrical characteristics (Fig. 31b).

Let us agree ENE, shown to Fig. 30a, b, e and their characteristics, to call 1 types ENE, and ENE, given in Fig. 30, c, d, f, and their characteristic - 2 types.

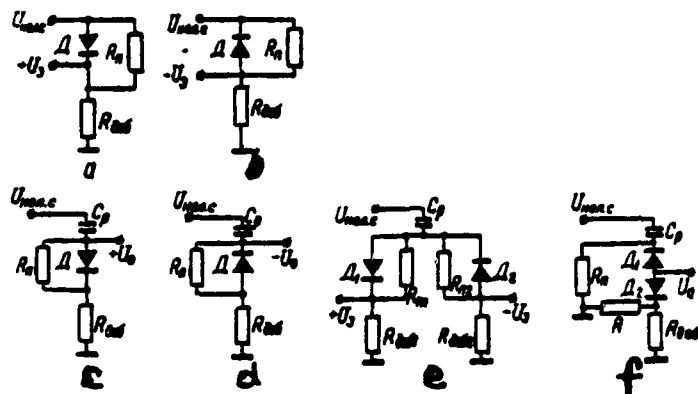


Fig. 30. The diagrams of the equivalent nonlinear elements/cells: a, b, c, d - asymmetric; e, f - symmetrical; a, b, e - first type; c, d, f - second type.

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In the diagram in Fig. 30f working diode for the positive signal is Δ_1 , and for negative Δ_2 . Diodes can be connected by the reversed polarity. Then the triggering voltage/stress must be negative. The methodology of the construction of dynamic characteristics of ENE is in detail examined in the work of the author [7, 8].

The form of characteristics of ENE can be changed, changing the values of that locking U_o or triggering U_o of voltages/stresses, and also resistors/resistances R_n and R_m , connected respectively in parallel and in series with nonlinear elements/cells [7].

It is obvious that ENE with the asymmetric characteristics one should apply for the formation of FAKh during the amplification of single-pole pulse signals, and ENE with the symmetrical characteristics - for the formation of FAKh during the amplification of bipolar pulse and harmonic signals.

Differential and average/mean conductivities of ENE with different approximations of volt-ampere characteristic can be calculated from the appropriate formulas for the differential and average/mean slope/transconductance of passage characteristic of UP.

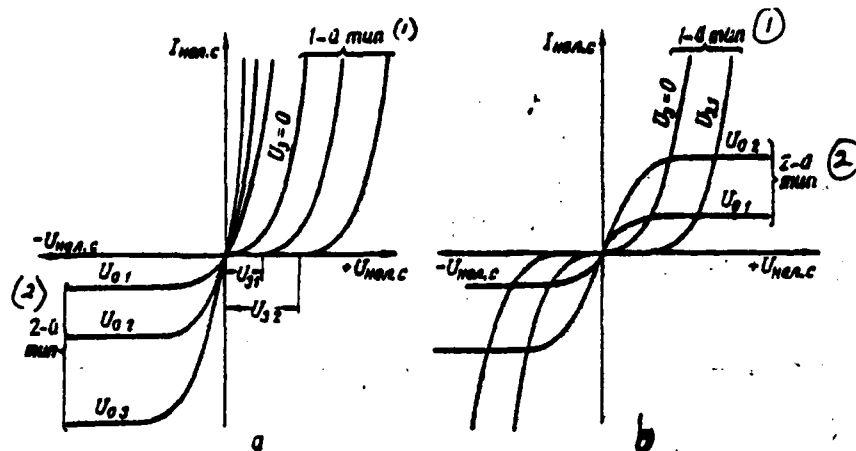


Fig. 31. Dynamic volt-ampere characteristics of ENE: a) asymmetric; b) symmetrical.

Key: (1). 1st type. (2). 2nd type.

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§3. Functional amplifiers with the nonlinear elements/cells in the load circuit.

As can be seen from expression (2-19), FAKh can be realized upon the inclusion of nonlinear element/cell into the load circuit both of aperiodic and selective electron-tube or transistor amplifier stage. However, FAKh in the broad dynamic band with the help of one cascade/stage in practice cannot be realized, it is possible to

realize it in the multistage amplifier. With series connection of cascades/stages and shunting of loads by nonlinear elements/cells [7] in selective FU it is expedient to apply cascades/stages with the single ducts/contours or the two-circuit filters, tuned also for one frequency.

The essence of realization of FAKh is identical for all amplifiers. Therefore for the analysis of FU we use a more overall equivalent diagram of resonance UK, given in Fig. 32.

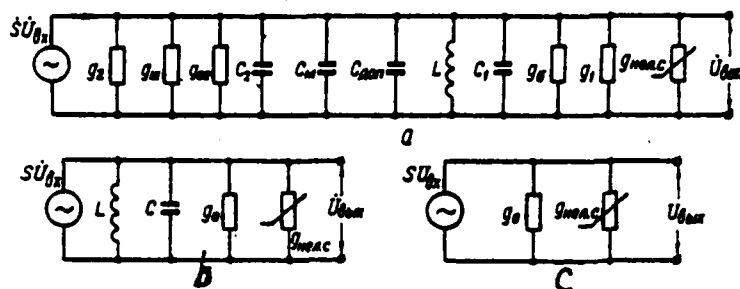


Fig. 32. Equivalent diagrams of resonance FU of voltage/stress with the load, shunted by the nonlinear element/cell: a, b) complete and convoluted the schematic of tuned amplifier; c) the convoluted schematic of the resonance and aperiodic amplifiers for the region of medium frequencies; g_1 and g_2 - input in output conductance UP; g_m - conductivity of anodic (or collector) resistor/resistance for the aperiodic amplifier or resisting the shunt for the tuned amplifier: g_{oe} - the equivalent conductivity of losses in the oscillatory circuit; g_0 - conductivity of the resistors/resistances of feed circuit in the case of the transistor amplifier: g_{neac} - the conductivity of nonlinear element/cell; C_1 and C_2 - input and output capacitances of UP; C_m - wiring capacitance; C_{don} - further capacity/capacitance; $Y = g_1 + g_m + g_0 + g_2$ - general/common/total load admittance without taking into account the conductivity of nonlinear element/cell; $C = C_1 + C_m + C_{don} + C_2$ - capacity/capacitance of oscillatory circuit.

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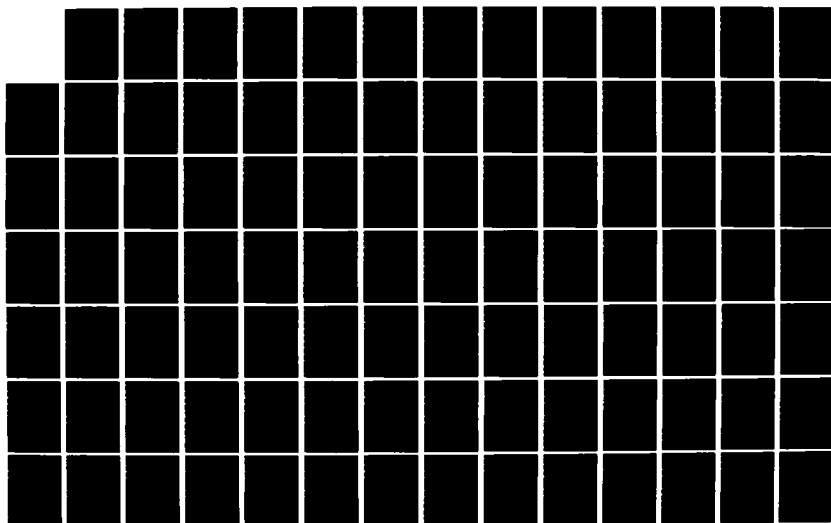
FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC
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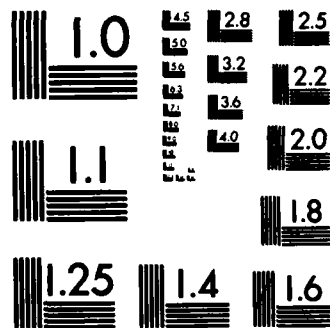
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General/common/total load admittance of the cascade/stage

$$g = g_0 + g_{\text{load}} \quad (2-96)$$

With the work of cascade/stage in the linear conditions

$$g_n = g_0 + g_{\text{load}, n} \quad (2-97)$$

FAKh of amplifier usually remove/take at the resonance frequency for the selective amplifiers or at the quasi-resonance frequency, which lies at the region of medium frequencies for the aperiodic amplifiers. Therefore for the analysis we use the equivalent diagram for the medium frequencies (Fig. 32c), suitable for the aperiodic and selective amplifiers. Let us consider some special features/peculiarities of selective transistorized amplifiers separately.

Using expression (2-19), let us find the requirements, by which it must satisfy nonlinear element/cell for the realization of required FAKh during the use of real UP.

LINEAR AMPLIFIER.

For the linear amplifier condition $z=x$ is satisfied. Then condition (2-19) can be registered in the form

$$g = g_0 \frac{g}{g_0} \quad (2-98)$$

since $g_n = g_0$.

For the linear amplifier in expression (2-98) it is possible to substitute the value of differential slope/transconductance.

Linear amplifier instrument. In this case

$$g = g_0 = \text{const}; \quad (2-99)$$

$$k_{\text{max}} = 0. \quad (2-100)$$

Amplifier instrument with the characteristic, approximated by the polynomial of the second power taking into account expression (2-39), for the positive pulse signal we obtain

$$g^{(-)} = g_0 \frac{g^{(+)}_0}{g_0} = g_0 [1 + (\eta - 1)q]. \quad (2-101)$$

Sign \leftarrow with g considers the rotation of the phase of amplified signal 180° . Conductivity of the nonlinear element/cell

$$g^{(-)}_{\text{max}} = g^{(-)} = g_0 = g_0 (\eta - 1)q. \quad (2-102)$$

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We find expression for volt-ampere characteristic of nonlinear element/cell. In this case we assume that conductivity $g_0 = \text{const}$ and does not depend on the value of signal. This assumption in many instances is correct for the vacuum-tube amplifiers, for the transistor amplifiers conductivity g_0 is changed as a result of a change in conductivity g_{ex} . It is possible to take into account a

change of conductivity g_{nx} subsequently and to correct the characteristic of nonlinear element/cell. Then

$$I_{n\alpha x}^{(-)} = \int g_{n\alpha x}^{(-)} dU_{n\alpha x} = \frac{g_0(\eta-1)}{U_{n\alpha, n}} \int U_{n\alpha} dU_{n\alpha x} + C.$$

Substituting $U_{n\alpha} = \frac{U_{n\alpha x}}{K_n}$, we obtain

$$I_{n\alpha x} = \frac{g_0(\eta-1)}{2K_n U_{n\alpha, n}} + C.$$

On the basis of initial conditions $I_{n\alpha x} = 0$ when $U_{n\alpha x} = 0$, integration constant $C=0$. Finally we obtain

$$I_{n\alpha x} = \frac{g_0^2(\eta-1)}{2S_n U_{n\alpha, n}} U_{n\alpha x}^2, \quad (2-103)$$

since

$$K_n = \frac{S_n}{g_0}.$$

The volt-ampere characteristic, described by expression (2-103), can be realized with the help of ENE, shown in Fig. 30b.

For the negative signal taking into account expression (2-39) we analogously obtain

$$g_{n\alpha x}^{(+)} = g^{(+)} - g_0 = -g_0(\eta-1)q, \quad (2-104)$$

i.e., in the case of amplifying the negative signal the conductivity of the nonlinear element/cell, connected in parallel to load admittance, it must be negative, which to in practice realize is very difficult (theoretically this can be done with the help of the tunnel diodes).

$$g_{\text{max}} \ll g_0$$

If in FU conditions Λ and $g_{\text{ax}} \ll g_0$ are satisfied, then should be found out the volt-ampere characteristic of nonlinear load and load fulfilled in the form of equivalent nonlinear element/cell. Then

$$g_{\text{uap}}^{(\mp)} = g_u \frac{S^{(\pm)}}{S_u} = g_u [1 \pm (\eta - 1) q]. \quad (2-105)$$

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Substituting the value $U_{\text{ax}} = \frac{U_{\text{max}}}{K_u}$ into expression (2-105) and integrating it, we obtain expression for the volt-ampere characteristic of the nonlinear load

$$I_{\text{uap}}^{(\mp)} = g_u \left(U_{\text{max}} \pm \frac{\eta - 1}{2K_u U_{\text{ax. u}}} U_{\text{max}}^2 \right), \quad (2-106)$$

where

$$g_u = \frac{S_u}{K_u}.$$

Volt-ampere characteristics, described by expression (2-106), are depicted in Fig. 33 (curve 1-1), it is possible to realize it with the help of ENE, shown in Fig. 30d.

Amplifier instrument with the characteristic, approximated by the polynomial of the third power. According to expressions (2-51) and (2-98) for the load admittance we obtain

$$g^{(\mp)} = g_u (1 - q^2). \quad (2-107)$$

After substitution $U_{\text{ax}} = \frac{U_{\text{max}}}{K_u}$ into expression (2-107) and integration finally we obtain

Amplifier instrument with exponential characteristic $i = I_0(e^{U/U_0} - 1)$.
Differential slope/transconductance for the bipolar signal

$$S^{(\pm)} = \gamma I_0 e^{\gamma(U_0 \pm U_{\max})} = S_0 e^{\pm \gamma U_{\max}}. \quad (2-111)$$

Then load admittance

$$g^{\pm} = g_0 \frac{S^{\pm}}{S_0} = g_0 e^{\mp \gamma U_{\max}}. \quad (2-112)$$

Current of the load

$$I_{\text{load}}^{(\pm)} = g_0 \int e^{\mp \gamma U_{\max}} dU_{\max} = \mp \frac{g_0}{\beta} e^{\mp \gamma U_{\max}} + C,$$

where

$$\beta = \frac{1}{K_0}.$$

We find integration constant from conditions $I_{\text{load}} = 0$ when $U_{\max} = 0$

$$C = \pm \frac{g_0}{\beta} e^{\mp \gamma U_{\max}}.$$

After the substitution of integration constant, we obtain

$$I_{\text{load}}^{\pm} = I_0 e^{\gamma U_0} (\mp e^{\mp \gamma U_{\max}} \pm 1), \quad (2-113)$$

since

$$g_0 K_0 = S_0 = \gamma I_0 e^{\gamma U_0}.$$

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Volt-ampere characteristic, described by expression (2-113), is analogous in form to characteristic (1-1) in Fig. 33 and can be realized with the help of ENE (Fig. 30d).

During the amplification of unipolar signal for expanding the dynamic range of the linear AKh of amplifier should be the operating

point on the passage characteristic of amplifier instrument displaced into the region of larger or smaller bias voltage $U_{a.}$ depending on signal polarity and type of UP (tube or transistor). For example, in the case of amplifying the positive signal it is necessary to increase initial negative bias voltage, if as UP is used vacuum tube, connected on the diagram "common cathode", or a transistor of type p-n-p, connected on the diagram "general/common/total emitter"; to reduce the initial positive bias voltage, if as UP is used a transistor of type n-p-n, connected on the diagram "general/common/total emitter".

NONLINEAR FUNCTIONAL AMPLIFIER.

In the nonlinear amplifiers for calculating the amplitude properties it is necessary to use an average/mean slope/transconductance of passage characteristic of UP.

For the synthesis of nonlinear FU we use condition (2-19)

$$g = g_n \frac{S}{S_n} \cdot \frac{S}{I} = g_n \frac{S}{S_n} \varphi(z).$$

Let us consider the possibility of realization of FAKh with various forms of passage characteristic of UP.

Linear amplifier instrument $S = S_n = \text{const.}$ In this case

$$g(z) = g_n \varphi(z). \quad (2-114)$$

Conductivity of nonlinear element/cell when $b = \text{var}(\downarrow)$

$$g_{\text{non}}^{(\pm)} = g^{(\pm)} - g_n = g_n [\varphi(z) - 1], \quad (2-115)$$

or

$$R_{\text{non}}^{(\pm)} = \frac{1}{g_{\text{non}}} = \frac{R_n}{\varphi(z) - 1}, \quad (2-116)$$

since $g_n = g_0$.

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In this case the required volt-ampere characteristic of the nonlinear element/cell

$$I_{\text{non}}^{(\pm)} = g_{\text{non}} U_{\text{max}} = g_0 U_{\text{non}} [\varphi(z) - 1], \quad (2-117)$$

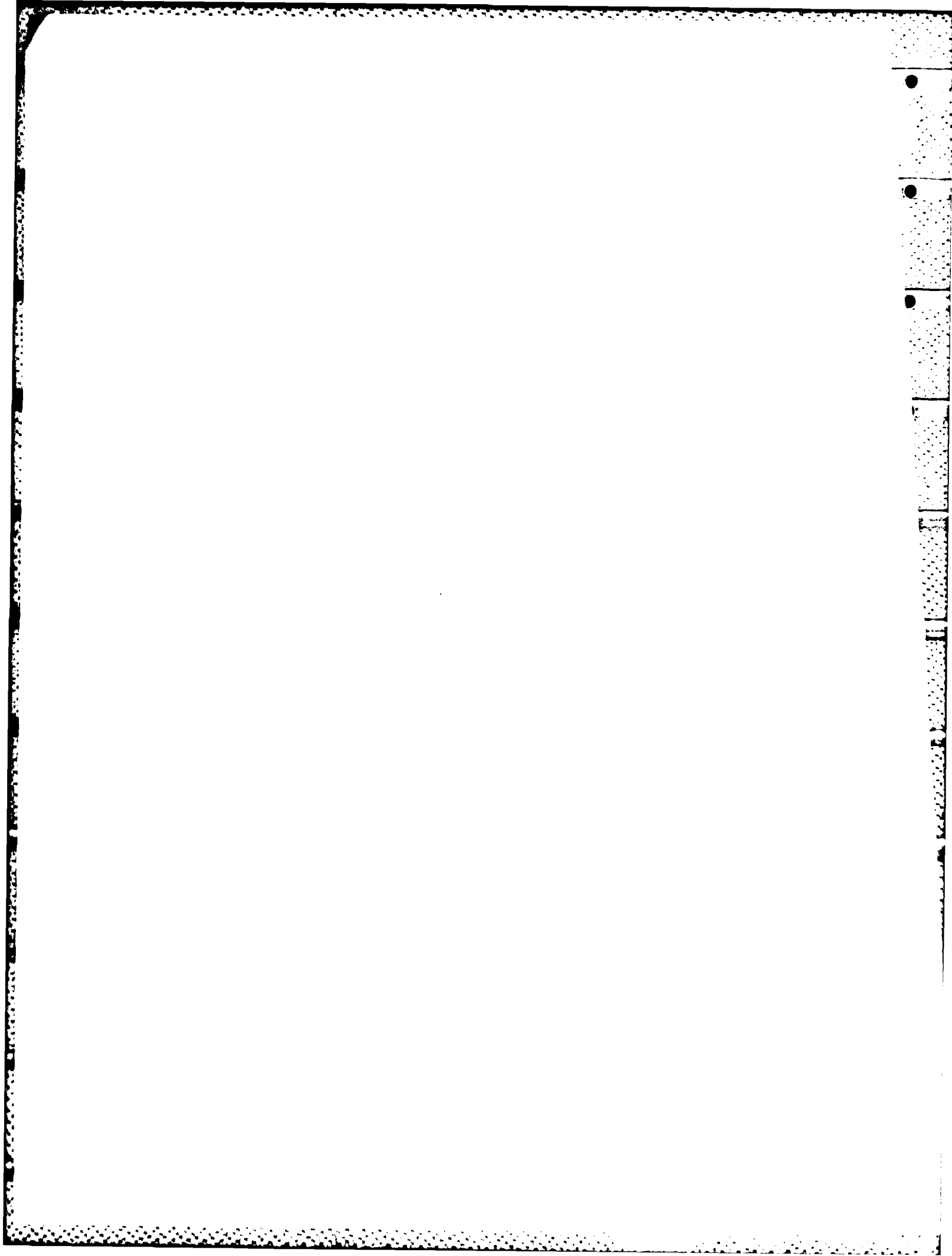
or

$$I_{\text{non}}^{(\pm)} = g_0 U_{\text{max}} \cdot n z [\varphi(z) - 1], \quad (2-118)$$

where $\varphi(z) = x/z$ - function, mathematical recording by which for different forms of FAKh is given in Table 5.

If one of the components of conductivity g_0 is changed with a change in voltage/stress U_{max} , then this can be taken into account in formula (2-118) during the calculation of the volt-ampere characteristic of nonlinear element/cell.

Expression (2-118) is general/common/total and suitable for calculating the nonlinear cascade/stage from FAKh when $b = \text{var}(\downarrow)$, which works in the linear, functional and quasi-linear modes/conditions.



For this it is necessary in expression (2-118) to substitute the value of z and $\varphi(z)$, corresponding to one of the operating modes.

In the case of FAKh, when differential coefficient grows over the dynamic range, i.e., $b = \text{var}(\uparrow)$, the conductivity of nonlinear element/cell $g_{\text{нел}}$, calculated by formula (2-115), is negative. Then it is necessary to calculate $g_{\text{нагр}}$ and current $I_{\text{нагр}}$ of nonlinear load. According to formula (2-114)

$$I_{\text{нагр}} = g_{\text{н}} U_{\text{вмх}} \frac{z}{z} = g_{\text{н}} U_{\text{вмх}} x(z), \quad (2-119)$$

where $x(z)$ - the standardized/normalized input voltage, expressed through the standardized/normalized output voltage/stress z .

Expressions for $x(z)$ are given in Table 1. In particular, for exponential amplifier $x(z) = a \ln z + 1$

$$I_{\text{нагр}}^{(\pm)} = g_{\text{н}} U_{\text{вмх}} (a \ln z + 1). \quad (2-120)$$

For the development/detection of character, and also for calculating the volt-ampere characteristics of nonlinear elements/cells and loads it is convenient to use the generalized standardized/normalized volt-ampere characteristics

$$H_{\text{нел}}^{(\pm)} = \frac{I_{\text{нел}}^{(\pm)}}{I_{\text{нагр. н}}^{(\pm)}} = z [\varphi(z) - 1] \quad (2-121)$$

when $b = \text{var}(\downarrow)$ and

$$H_{\text{нагр}}^{(\pm)} = \frac{I_{\text{нагр}}^{(\pm)}}{I_{\text{нагр. н}}^{(\pm)}} = x(z) \quad (2-122)$$

when $b = \text{var}(\uparrow)$, where $I_{\text{нагр. н}} = g_{\text{н}} U_{\text{вмх. н}}$ - current, flowing through the load with the output voltage/stress, with which it begins with FAKh of amplifier.

The normalized characteristics for three types of FAKh with the amplification of a single-polar signal are presented in Fig. 34. For LAKh it is accepted $d=10$. During the amplification of the

bipolar signal of characteristic they must be symmetrical relative to the origin of coordinates and take the form of the characteristics, given in Fig. 31b,.

As can be seen from Fig. 34, FAKh when $b = \text{var}(\downarrow)$ it is possible to realize with the help of ENE of the first type, in this case to the nonlinear element/cell must be given cutoff voltage, equal to $U_0 = U_{\text{BML. N}^2\text{N}} = U_{\text{BML. N}}$ since $z_N = 1$. With $b = \text{var}(\uparrow)$ FAKh it is possible to realize with the aid of ENE of the second type, in this case to the nonlinear element/cell must be given triggering voltage/stress, which ensures the linearity of dynamic volt-ampere characteristic to voltage/stress $U_{\text{HOL}} = U_{\text{BML. N}}$.

Let us consider the possibility of the realization of characteristic 1 in Fig. 34 with the help of the diagram in Fig. 30a. It is obvious that the nonlinear section of characteristic from z_N to z_{N1} , that corresponds to the work of cascade/stage in the logarithmic mode/conditions, must ensure the nonlinear element/cell (diode). The approximating coefficients of volt-ampere characteristic of diode can be found from two points, for example 1 and 2. For points 1 and 2

$$I_{\text{HOL}1} = U_{\text{BML. N}^2\text{N}} \left(\frac{e^{z_1} - 1}{z_1} \right) = i_0 e^{U_{\text{BML. N}^2\text{N}}},$$

$$I_{\text{HOL}2} = I_{\text{HOL. N}} = i_0 e^{U_{\text{BML. N}^2\text{N}}}.$$

Solving equations relatively γ , we have

$$\gamma = \frac{\ln A}{U_{\text{BML. N}} (z_N - z_1)}, \quad (2-123)$$

where

$$A = \frac{I_{\text{нел. н}}}{I_{\text{нел. 1}}}$$

The greatest accuracy of approximation is obtained, if we as point 2 take the point, which corresponds z_n . In this case

$$I_0 = \frac{I_{\text{нел. 2}}}{\exp \gamma U_{\text{вмк. н}}^2 \text{ н}}. \quad (2-124)$$

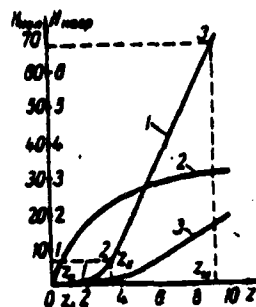


Fig. 34. The standardized/normalized volt-ampere characteristics of ENE in the implementation of different FAKh: 1 - logarithmic; 2 - exponential; 3 - quadratic.

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Section 2-3 in essence forms/shapes supplementary resistor/resistance $R_{\text{дог}}$, connected in series with the diode. The value of resistor/resistance $R_{\text{дог}}$ can be found as follows. For point 3, which corresponds to the end/lead of the quasi-linear section of characteristic (to maximum output voltage/stress of nonlinear cascade/stage),

$$U_{\text{вхл. н}} = U_{\text{нел. н}} + I_{\text{нел. н}} (R_{\text{дог}} + R_{\text{д}}),$$

where

$$U_{\text{нел. н}} = \gamma \ln \frac{I_{\text{нел. н}}}{i_0}; \quad I_{\text{нел. н}} = U_{\text{вхл. н}} g_{\text{нл}} [\varphi(z) - 1];$$

$R_{\text{д}}$ — the strength of materials of diode, the component of the unit of ohms. Whence

$$R_{\text{дог}} = \frac{U_{\text{вхл. н}} - \gamma \ln \frac{I_{\text{нел. н}}}{i_0}}{I_{\text{нел. н}}}. \quad (2-125)$$

Resistor/resistance $R_{\text{дог}}$ is, as a rule, small and it virtually does not affect the form of the section of 1-2 characteristics.

Analogously it is possible to calculate elements/cells of ENE of the second type.

Amplifier instrument with the nonlinear passage characteristic.
Conductivity of the nonlinear element/cell

$$g_{\text{non}} = g_n [\Phi(S) \Phi(z) - 1]. \quad (2-126)$$

For finding the volt-ampere characteristics of the nonlinear ones of element/cell and load we will use equation (2-19). On the basis of expressions (2-19), (2-121) and (2-122), we obtain

$$H(I_{\text{non}}^{\pm}) = z \left[\frac{S^{\pm}}{S_n} \varphi(z) - 1 \right] = z [\Phi(S) \varphi(z) - 1] \quad (2-127)$$

with $b = \text{var}(\downarrow)$,

$$H(I_{\text{non}}^{\pm}) = \frac{S^{\pm}}{S_n} x(z) = \Phi(S) x(z) \quad (2-128)$$

with $b = \text{var}(\uparrow)$.

Expressions for standardized/normalized characteristics (2-127) with $b = \text{var}(\downarrow)$ are given in Tables 7 and for characteristics (2-128)

with $b = \text{var}(\uparrow)$ Table 8. In this case are accepted the following

designations: $p = \frac{U_{\text{ex. n}}}{U_{\text{rs}}}$ — for the polynomial of the third power;

$p = \frac{U_{\text{ex. n}}}{U_{\text{ex. n}}}$ — for the polynomial of the second power; $\beta = qU_{\text{ex. n}}$; $m = \frac{U_{\text{ex. n}}}{U_{\text{ex. n}}}$ — for the hyperbolic tangent; $\xi = U_{\text{ex. n}}$ — for the exponential curve.

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The schematic diagram of four-stage logarithmic amplifier on the tubes with the single ducts/contours, shunted by nonlinear elements/cells, is given in work [7]. Amplifier has the following parameters: resonance frequency $f_0=20$ MHz; amplification factor at the work in linear conditions $K_0=4.1 \cdot 10^3$ or $K_{0(0.01)}=72.7$ dB; passband with the work in the linear conditions $\Pi=0.4$ MHz. Dynamic range of LAKh is $D=70$ dB, which indicates the strictly successive work of cascades/stages. LAKh begins with input voltage $U_{\text{вх}} = 2.5 \cdot 10^{-5}$ V. Relative accuracy of LAKh in the voltage/stress in entire range $\delta U(2-3)\%$. Slope/transconductance of LAKh $\sigma=18$ mV/dB.

The schematic diagram of single-stage amplifier with EAKh with second type asymmetric nonlinear element/cell in the anode circuit is given in work [16]. Dynamic range of EAKh on the input voltage $d=20$ dB, on the output $-d_{\text{вых}}=40$ dB.

Table 7.

 $H(I_{\text{нел}}^{\pm})$ при $b = \text{var}(\downarrow)$

Аппроксимация (1)	Тип сигнала (2)	
	Импульсный (3)	Гармонический (4)
Полином 2-й степени (5)	$\frac{I_{\text{нел}}^{\pm}}{I_{\text{нагр. н}}} = \left\{ \left[1 \mp \frac{(\gamma-1)xp}{2} \right] \frac{x}{s} - 1 \right\} z$	0
Полином 3-й степени (5)	$\frac{I_{\text{нел}}}{I_{\text{нагр. н}}} = z \left[\left(1 - \frac{x^2 p^2}{3} \right) \frac{x}{s} - 1 \right]$	$\frac{I_{\text{нел}}}{I_{\text{нагр. н}}} = z \left[\left(1 - \frac{x^2 p^2}{4} \right) \frac{x}{s} - 1 \right]$
Гиперболический тангенс (6)	$\frac{I_{\text{нел}}}{I_{\text{нагр. н}}} = -z \left(\frac{\text{sh } \beta m \text{ ch } \beta}{U_{\text{нел. н}}^2 \text{ ch } \beta (1 \pm m)} - 1 \right)$	-
Экспонента (7)	$\frac{I_{\text{нел}}}{I_{\text{нагр. н}}} = z \left(\frac{\pm e^{Lx} \mp 1}{Ls} - 1 \right)$	$\frac{I_{\text{нел}}}{I_{\text{нагр. н}}} = z \left\{ \left[1 + \frac{\gamma^2 U^2}{2s} + \frac{\gamma^4 U^4}{3 \cdot 2s} + \dots \right] \frac{x}{s} - 1 \right\}$

Key: (a). with. (1). Approximation. (2). Type of signal. (3). Pulse. (4). Harmonic. (5). Polynomial of 2nd degree. (6). Hyperbolic tangent. (7). Exponential curve.

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As can be seen from expression (2-19), general/common/total load admittance of FU with an increase in the signal must be changed. This leads to a change in the frequency, phase and transient responses of

FU with an increase in the signal. It is necessary to note that FU are nonlinear amplifiers. Therefore about the passband, the phase, the upper and lower cut-off frequencies it is possible to speak conditionally for steady-state mode/conditions under the influence of continuous harmonic signal. For the pulse selective and aperiodic amplifiers it is necessary to examine transient process in the nonlinear system and to determine the distortions of front and flat/plane apex/vertex of the reinforced pulse signal.

Absolute and relative changes of the passband, upper and lower cut-off frequencies, and also phase shift from the dynamic range can be calculated, substituting in the known formulas of the enumerated parameters of the value of load admittance for this signal level.

Table 8.

$H(I_{\text{нагр}}^{\pm})$ при $b = \text{var}(\uparrow)$		
Аппроксимация (1)	(2) Тип сигнала	
	Импульсный (3)	Гармонический (4)
Полном 2-й степени (5)	$\frac{I_{\text{нагр}}^{\pm}}{I_{\text{нагр. н}}} = \left[1 \mp \frac{(\eta-1)xp}{2} \right] x$	$\frac{I_{\text{нагр}}}{I_{\text{нагр. н}}} = x$
Полном 3-й степени (5)	$\frac{I_{\text{нагр}}^{\pm}}{I_{\text{нагр. н}}} = \left(1 - \frac{x^2 p^2}{3} \right) x$	$\frac{I_{\text{нагр}}}{I_{\text{нагр. н}}} = \left(1 - \frac{x^2 p^2}{4} \right) x$
Гиперболический тангенс (6)	$\frac{I_{\text{нагр}}^{\pm}}{I_{\text{нагр. н}}} = U_{\text{вх. н}} \frac{\text{sh } \beta m \text{ ch } \beta}{\text{ch } \beta (1 \pm m)}$	—
Экспонента (7)	$\frac{I_{\text{нагр}}^{\pm}}{I_{\text{нагр. н}}} = \frac{\pm e^{ix} \mp 1}{\epsilon}$	$\frac{I_{\text{нагр}}}{I_{\text{нагр. н}}} = \left(1 + \frac{\gamma^2 U^2}{2^2} + \frac{\gamma^4 U^4}{3 \cdot 2^4} \dots \right) x$

Key: (1). Approximation. (2). Type of signal. (3). Pulse. (4). By harmonic. (5). Polynomial of 2nd degree. (6). Hyperbolic tangent. (7). Exponential curve.

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It is easy to show that for FU with $b = \text{var}(\uparrow)$ with an increase in the number of nonlinear cascades/stages an absolute change in the phase in the dynamic range grows, and a relative change in the

passband is reduced. Consequently, requirements for the selection of a number of nonlinear cascades/stages from the point of view of the smallest changes in phase and passband are contradictory.

FU WITH NONLINEAR DIVIDERS.

A change in the frequency parameters of FU in the dynamic range can be decreased, if we decrease the range of a change in the shunting effect of nonlinear element/cell in a some manner. For this nonlinear element/cell must be included/connected consecutively/serially with the sufficiently high linear resistor/resistance (Fig. 35a, b, c) or by load (Fig. 35d), and the voltage of signal on the input of the following cascade/stage removed/taken from the formed divider. Let us agree the chain/network of series-connected resistors/resistances R and R_{non} to call a nonlinear divider of the 1st type (Fig. 35a) and the 2nd type (Fig. 35b).

In the given diagrams of selective FU the frequency, at which the transmission factor of cascade/stage is maximum, differs from the resonance frequency of oscillatory circuit. Because of this the analysis of selective FU must be carried out taking into account all reactances of oscillatory system and it is not possible to obtain overall equivalent diagram for the aperiodic ones and from the

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selective amplifiers, that considers only active component load admittances.

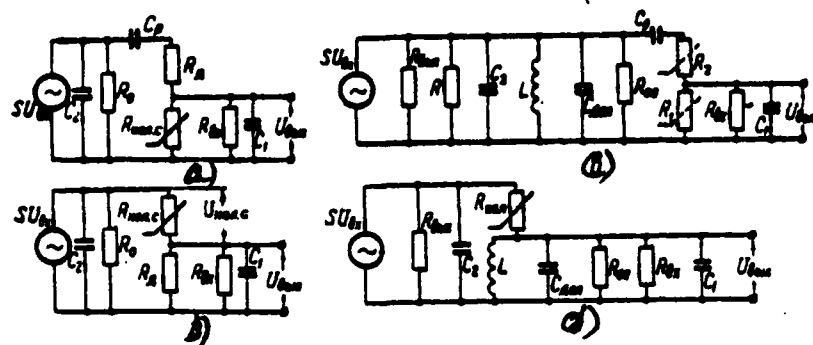


Fig. 35. Equivalent diagrams of FU with the nonlinear dividers for FU: a, b) aperiodic; c, d) resonance.

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Therefore the analysis of aperiodic and selective amplifiers must be carried out separately.

Aperiodic FU. In the region of medium frequencies is disregarded the effect of stray capacitances. Then transmission factor FU with 1st type nonlinear divider (Fig. 35a)

$$K = \frac{SR_0 R_{\text{нел}} R_{\text{нх}}}{R_{\text{нел}} (R_0 + R_{\text{н}} + R_{\text{нх}}) + R_{\text{нх}} (R_0 + R_{\text{н}})}, \quad (2-129)$$

where

$$\frac{1}{R_0} = \frac{1}{R_{\text{нел}}} + \frac{1}{R_{\text{н}}}$$

Taking into account condition (2-19), we obtain expression for

the resistor/resistance

$$R_{\text{non.c}} = \frac{1}{s_{\text{non}}} = \frac{R_{\text{bx}} (R_0 + R_{\text{R}})}{[\varphi(s) \Phi(S) + B - 1] (R_0 + R_{\text{R}} + R_{\text{bx}})}, \quad (2-130)$$

where $B = \frac{R_{\text{bx}} (R_0 + R_{\text{R}})}{R_{\text{non. n}}}$; $R_{\text{non. n}}$ — initial resistor/resistance of nonlinear element/cell with the work of FU in the linear conditions.

For FAKh with $b=\text{var } (\uparrow)$ the resistor/resistance

$$R_{\text{non. n}} = \frac{K_{\text{R}} R_{\text{bx}} (R_0 + R_{\text{R}})}{s_{\text{R}} R_0 R_{\text{bx}} - K_{\text{R}} (R_0 + R_{\text{R}} + R_{\text{bx}})}.$$

For FAKh with $b=\text{var } (\downarrow)$ resistor/resistance $R_{\text{non. n}} = \infty$ and coefficient $B=0$. In this case

$$R_{\text{non.c}} = \frac{1}{s_{\text{non.c}}} = \frac{R_{\text{bx}} (R_0 + R_{\text{R}})}{[\varphi(s) \Phi(S) - 1] (R_0 + R_{\text{R}} + R_{\text{bx}})}. \quad (2-131)$$

From expressions (2-157) and (2-159) it is evident that the resistor/resistance of nonlinear element/cell with the rotor of signal with $b=\text{var } (\uparrow)$ must grow, with $b=\text{var } (\downarrow)$ — be reduced.

For the vacuum-tube amplifiers inequality $R_{\text{bx}} \gg R_0 + R_{\text{R}}$ usually is fulfilled. Then expressions (2-130) and (2-131) are simplified

$$R_{\text{non.c}} = \frac{R_0}{\varphi(s) \Phi(S) + B - 1} \left(1 + \frac{R_{\text{R}}}{R_0} \right); \quad (2-132)$$

$$R_{\text{non.c}} = \frac{R_0}{\varphi(s) \Phi(S) - 1} \left(1 - \frac{R_{\text{R}}}{R_0} \right). \quad (2-133)$$

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The dynamic volt-ampere characteristic of nonlinear element/cell can be calculated from formula (2-117). With $b=\text{var}$ (†) is required nonlinear element/cell with 1st type characteristic with $b=\text{var}$ (†) — with 2nd type characteristic.

For amplifier stage with 2nd type nonlinear divider (Fig. 35b) the transmission factor in the region of the medium frequencies

$$K = \frac{S R_0 R_n R_{nx}}{(R_n + R_{nx})(R_0 + R_{nx}) + R_n R_{nx}}. \quad (2-134)$$

The resistor/resistance of nonlinear element/cell must be changed according to the law

$$R_{nx, c} = \frac{1}{\epsilon_{nx}} = \varphi(z) \Phi(S) (R_0 + R_n + R_{nx, n}) - R_0 - R_n, \quad (2-135)$$

where

$$R_{nx, n} = \frac{S_n R_0 R_n R_{nx} - K_n [R_0 (R_n + R_{nx}) + R_n R_{nx}]}{K_n (R_n + R_{nx})}.$$

For vacuum-tube amplifiers $R_{nx} \gg R_n$. Then

$$\begin{aligned} R_{nx, n} &= \frac{S_n R_0 R_n}{K_n} - R_0 - R_n; \\ R_{nx, c} &= \varphi(z) \Phi(S) \frac{S_n R_0 R_n}{K_n} - R_0 - R_n. \end{aligned} \quad (2-136)$$

For the selection of nonlinear element/cell during the calculation of FU it is necessary to have dependence $R_{nx, c} = f(U_{nx, c})$ or

dynamic volt-ampere characteristic, which can be found from the relationship/ratio

$$z_{\text{non. c}} = \frac{U_{\text{non. c}}}{U_{\text{max. c}}} = z \frac{R_{\text{non. c}}}{R'_x}. \quad (2-137)$$

Whence

$$U_{\text{non. c}} = U_{\text{max. c}} z \frac{R_{\text{non. c}}}{R'_x}. \quad (2-138)$$

In this case the signal current, flowing through the nonlinear element/cell,

$$I_{\text{non. c}} = \frac{U_{\text{non. c}}}{R_{\text{non. c}}} = \frac{U_{\text{max. c}}}{R'_x}, \quad (2-139)$$

where

$$R'_x = \frac{R_x R_{\text{ex}}}{R_x + R_{\text{ex}}}.$$

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The dynamic volt-ampere characteristic of nonlinear element/cell is constructed with the joint graphical solution of equations (2-138) and (2-139).

Analyzing expressions (2-135), (2-136) and (2-139), it is possible to draw the conclusion that for the realization of FAX with $b=\text{var } (\downarrow)$ is required the nonlinear element/cell with 1st type characteristic, with $b=\text{var } (\uparrow)$ - the 2nd type.

For obtaining the broader dynamic band FAKh with $b=\text{var}$ (\uparrow) initial factor of amplification K_u with the work of FU in the linear conditions must be chosen as smaller as possible. In this case value $R_{\text{uez. c}}$ is small for 1st type nonlinear divider and, on the contrary, large - for 2nd type nonlinear divider.

Transmission factor of FU, made on the diagrams in Fig. 35a, b in the region of highest frequencies,

$$K_u = \frac{SRR_1}{\sqrt{A^2 + B^2}}; \quad (2-140)$$

$$A = f - \omega^2 h; \quad B = \omega e; \quad f = R + R_1 + R_2; \quad h = C_2 C_1 R R_1 R_2;$$

$$e = C_1 R_1 R + C_1 R_1 R_2 + C_2 R (R_1 + R_2).$$

$$R_1 = R_{\text{uez. c}}; \quad R_2 = R_R \text{ (puc. 35, a)}; \quad R_1 = R_R; \quad R_2 = R_{\text{uez. c}} \text{ (puc. 35, b)}.$$

Key: (1). Fig.

Upper cut-off frequency f_{maxo} on which K is reduced $\sqrt{2}$ times,

$$f_{\text{maxo}} = \frac{1}{2\pi} \sqrt{\frac{-(e^2 - 2/h) \pm \sqrt{(e^2 - 2/h)^2 + 4/f^2 h^2}}{2h^2}}. \quad (2-141)$$

From expression (2-141) it is evident that f_{maxo} depends on signal level (resistors/resistances (R_1 and R_2)) and with $b=\text{var}$ (\uparrow) frequency f_{maxo} with an increase in the signal increases, and with $b=\text{var}$ (\downarrow)— it is reduced. As it is not difficult to see general/common/total stray capacitance of diagram, which shunts load, from the diagrams in Fig. 35a, b with $R_1 \gg R_2$, is equal to $C_2 + C_1$;

with $R_1 \ll R_2$, capacity/capacitance $C_0 \rightarrow C_1$. Then a maximum relative change of frequency f_{max} as a result of changing the resistors/resistances of R_1 and R_2 does not exceed the value

$$n = \frac{f'_{\text{max}}}{f_{\text{max}}} < \frac{C_2 + C_1}{C_2}. \quad (2-142)$$

With $C_2 = C_1$, relation $n \leq 2$.

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Lower cut-off frequency f_{min} for the diagrams in Fig. 35a, b

$$f_{\text{min}} = \frac{1}{2\pi C_p (R' + R'_1 + R_2)}, \quad (2-143)$$

где $\frac{1}{R'} = \frac{1}{R} + \frac{1}{R_{\text{max}}}$; $\frac{1}{R'_1} = \frac{1}{R_1} + \frac{1}{R_{\text{mx}}}$.

With $b = \text{var} (\downarrow)$ frequency f_{min} increases, with $b = \text{var} (\uparrow)$ —it is reduced. Let us consider the worst case, when f_{min} increases. Maximum value f'_{min} with $R_1 \rightarrow 0$. Then the relation

$$m = \frac{f'_{\text{min}}}{f_{\text{min}}} < \frac{R' + R'_1 + R_2}{R' + R_2} = 1 + \frac{R'_1}{R' + R_2}. \quad (2-144)$$

With $R_1 = R' + R_2$, relation $m \leq 2$.

Thus, in the aperiodic fie with the load, shunted by nonlinear divider, by the selection of the corresponding value of the linear resistor/resistance of divider R_2 it is possible to considerably decrease the instability of cut-off frequencies over the dynamic

range.

Work [7] gives the schematic diagrams of the bipolar (Fig. 52) and unipolar (Fig. 58) video amplifiers, in which LAKh is obtained with the shunting of plate loads by the nonlinear dividers respectively of the 1st and 2nd types. Bipolar amplifier has following data: $K_0 = 10^3$; upper cut-off frequency with the work in linear conditions $f_{max} = 3 \cdot 10^6$ Hz; $U_{ex. n} = 1.3 \cdot 10^{-4}$ V; $D \approx 60$ dB; relative accuracy of LAKh in entire dynamic range $\delta U \approx 3\%$; the slope/transconductance of LAKh $\sigma \approx 13$ mV/dB.

Unipolar amplifier has following data: $K_0 = 10$; $f_{max} = 2.5$ MHz; $\dot{U}_{ex. n} \approx 10^{-3}$ in; $d \approx 75-80$ dB; relative accuracy of LAKh $\delta U = 3-4\%$; $\sigma \approx (9-10)$ mV/dB; accuracy in the slope/transconductance $\delta \sigma \approx (15-20)\%$.

The results of detailed research of the diagrams indicated are given in the same work.

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Resonance FU. The complex transmission factor of the amplifier, made on the diagram in Fig. 35c,

$$K = S\omega LR_1 \left[\frac{B}{A^2 + B^2} - j \frac{A}{A^2 + B^2} \right], \quad (2-145)$$

where

$$\begin{aligned} A &= \omega^2 [LC(R_1 + R_2) + LC_1R_1 + C_1C_2R_1R_2] - R_1 - R_2; \\ B &= \omega [L + Cr(R_1 + R_2) + C_1R_1R_2] - \omega^3 LCC_1R_1R_2; \end{aligned}$$

$r = r_1 + r_2$; r_1 - resistor/resistance of losses in the inductive branch of oscillatory circuit; r_2 - resistor/resistance of losses in the capacitive branch of duct/contour; $C = C_1 + C_{\text{con}} + C_{\text{m}}$ - capacity/capacitance of single oscillatory circuit.

Modulus/module of transmission factor

$$K = \frac{S\omega LR_1}{\sqrt{A^2 + B^2}}. \quad (2-146)$$

Phase of the amplifier

$$\varphi = -\arctg \frac{A}{B}. \quad (2-147)$$

Equalizing to zero imaginary parts of expression (2-15), we find the resonance frequency of system through the output

$$\omega_p = \omega_0 \sqrt{\frac{1+p}{1+p+n+mrR_2}}, \quad (2-148)$$

where $\omega_0 = 1/\sqrt{LC}$ - angular frequency of single oscillatory circuit without taking into account capacitor C_1 ; $p = R_2/R_1$; $n = C_1/C$; $m = C_1/L$.

From equations (2-147) and (2-148) it is evident that the resonance frequency and phase are changed with a change in

resistors/resistances of R_1 and R_2 . In the wideband amplifiers this change can be observed in the limits of passband. The changes of the frequency during the phases are less, the less the coefficients n and m , i.e., the less the capacitor C_1 and the greater the initial value p . Let us consider two extreme values:

$$1. R_1 = 0; \omega_p' = \omega_0 \sqrt{\frac{1}{1+n}}; 2. R_1 = 0; \omega_p'' = \omega_0.$$

Then a maximally possible change in the resonance frequency of the system

$$\gamma = \frac{\omega_p''}{\omega_p'} = \sqrt{1+n}.$$

Whence at the given value γ the capacity/capacitance of oscillatory circuit must be equal to

$$C = \frac{C_1}{\gamma^2 - 1}. \quad (2-149)$$

Real relative change in the resonance frequency

$$\frac{\omega_{p,u}}{\omega_p} = \sqrt{\frac{(1+p_u)(1+p+n+mR_u)}{(1+p)(1+p_u+n+mR_{u,u})}}. \quad (2-150)$$

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Let us consider two limiting cases of changing the phase.

1. $R_1 = 0$, which corresponds to work of FU in linear conditions.

In this case the phase of the amplifier

$$\varphi = -\arctg \frac{Y'}{Y_0},$$

where d'_0 — circuit damping taking into account by-passing of resistor/resistance of R_1 . For the vacuum-tube amplifiers usually $R_1 > R$. Then $d'_0 = d_0$. In this case

$$\nu' = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}.$$

2. $R_1 = 0$. Usually $R_1 \gg R$. Then d_0 — attenuation of single duct/contour with the losses. In this case

$$\nu' = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}.$$

In this case the phase drift will be observed in essence due to a change in the resonance frequency of oscillatory system.

For the amplifier Fig. 35d complex transmission factor neglecting of resistor/resistance R_{max}

$$K = S\omega L \left[\frac{B}{A^2 + B^2} - j \frac{A}{A^2 + B^2} \right]. \quad (2-151)$$

Here

$$\begin{aligned} A &= \omega^2 (LC + CC_1 r R_{max} + C_1 L) - 1; \\ B &= \omega C r - \omega^2 L C C_1 R_{max} + \omega C_1 R_{max}. \end{aligned}$$

where $C = C_{max} + C_m + C_1$ — capacity/capacitance of oscillatory circuit without taking into account capacity/capacitance C_1 ; r — resistance

of losses in the duct/contour taking into account resistor/resistance R_{ex} .

Modulus/module

$$K = \frac{S \omega L}{\sqrt{A^2 + B^2}}. \quad (2-152)$$

Phase is determined from formula (2-177).

The resonance frequency of the system

$$\omega_p = \frac{\omega_0}{\sqrt{1 + n' + m' r R_{\text{ex}}}}, \quad (2-153)$$

where $n' = C_2/C$, $m' = C_2/L$.

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Change in the resonance frequency in the range

$$\frac{\omega_{p. \text{н}}}{\omega_p} = \sqrt{\frac{1 + n' + m' r R_{\text{ex. н}}}{1 + n' + m' r R_{\text{ex. п}}}}. \quad (2-154)$$

Relation $\frac{\omega_{p. \text{н}}}{\omega_p} \rightarrow 1$ with $C \gg C_2$, which can be made in the narrow-band system.

Thus, nonlinear dividers can be applied only in the narrow-band tuned amplifiers. In the wide-band amplifiers the drift of resonance

frequency and a change of the phase in the dynamic range can reach the significant magnitudes.

WITH TWO-CIRCUIT FILTER BAND FUNCTIONAL AMPLIFIER.

The filter, which consists of two coupled circuits, tuned for one frequency, is simplest. The equivalent diagram of nonlinear cascade/stage with the two-circuit filter is shown in Fig. 36.

For obtaining the required by AKh cascade/stage it is possible to shunt with nonlinear elements/cells one or both coupled circuits. With shunting of both ducts/contours a number of nonlinear elements/cells grows doubly, which is economically disadvantageous. Furthermore, the types of the nonlinear elements/cells, which shunt both ducts/contours, in the general case must be different, since the voltage of signal, applied to secondary circuit U_{n2} , in the connection/communication, different from the critical, is not equal to voltage/stress U_{n1} , applied to the first duct/contour.

In the vacuum-tube amplifiers to the first duct/contour large direct/constant voltage is applied, in consequence of which upon the inclusion of nonlinear element/cell into the first duct/contour isolating capacitors with the increased operating voltage/stress will be required.

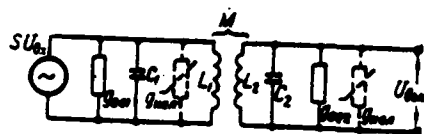


Fig. 36. The equivalent diagram of nonlinear cascade/stage with the two-circuit filter: $g_{01} = g_{001} + g_{001} + g_{01}$ - the worsened/impaired resonance conductivity of the first (output) duct/contour; $g_{02} = g_{002} + g_{002} + g_{02}$ - worsened/impaired resonance conductivity of the second (input) duct/contour; g_{01}, g_{02} - conductivity of further shunts.

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Therefore nonlinear element/cell to more expediently include in the secondary circuit. For the transistor amplifiers this does not have a value.

The connection/communication between the ducts/contours can be most varied. In the inductive coupling coupling coefficient

$$k_{12} = \frac{M}{\sqrt{L_1 L_2}}, \quad (2-155)$$

where M - mutual induction between the ducts/contours.

For the diagram in question complex transmission factor

$$K = \frac{j\omega S M}{\left[1 + \frac{j\omega}{\omega_0} b_1 + \left(\frac{j\omega}{\omega_0}\right)^2\right] \left[1 + \frac{j\omega}{\omega_0} b_{20} + \left(\frac{j\omega}{\omega_0}\right)^2\right] - \left(\frac{j\omega}{\omega_0}\right) b_1 b_{20} \beta^2 \left(b_1 + \frac{j\omega}{\omega_0}\right) \left(b_{20} + \frac{j\omega}{\omega_0}\right)}, \quad (2-156)$$

where $b_1 = \frac{g_{01}}{\omega_0 C_1}$; $b_{20} = \frac{g_{20}}{\omega_0 C_2}$; $g_{20} = g_{02} + g_{\text{nonl}}$;

$$C_1 = C_{01} + C_{M1} + C_{\text{nonl1}}; C_2 = C_{02} + C_{M2} + C_{\text{nonl2}}; \beta = \frac{k_{\text{on}}}{\sqrt{b_1 b_{20}}} -$$

the degree of the connection/communication between the ducts/contours; $\omega_0 = \frac{1}{\sqrt{L_1 C_1}} = \frac{1}{\sqrt{L_2 C_2}}$ - resonance frequency.

A change in the time lag of signal with a change in the value of the resistor/resistance of nonlinear element/cell is one of the deficiencies/lacks in amplifier cascades/stages with the load, shunted by nonlinear element/cell. It proves to be that during the appropriate selection of the sizes of the elements of oscillatory circuits and connection/communication between the ducts/contours this change in the time lag of signal can be considerably decreased.

Let us find the condition of the minimum of a change in the time lag with the shunting by the nonlinear element/cell of the first duct/contour. Let us determine the expression of the phase response of nonlinear cascade/stage $\phi(\omega)$. For simplification in the $K(j\omega)$, linings/calculations we will use the expression of the abbreviated

factor of amplification $K(\Omega)$, the methodology of determination of which is given in work [29]. In this case accordingly [29] in expression (2-156) let us take $\omega = \omega_0 + \Omega$, where Ω - small detuning.

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Taking into account the small first-order quantities: relative detuning $\frac{\Omega}{\omega_0}$, of attenuation δ_1 and δ_2 , and also the small second-order quantity, and throwing/rejecting all values, the order of smallness of which above second, we obtain

$$K(\Omega) = K_0 \frac{-\delta_1 \delta_2 (1 + \beta^2)}{4 \left(\frac{\Omega}{\omega_0} \right)^2 + 2(\delta_1 + \delta_2) \frac{\Omega}{\omega_0} + \delta_1 \delta_2 (1 + \beta^2)} \quad (2-157)$$

where K_0 - factor of amplification of cascade/stage in the presence of the resonance

$$K_0 = \frac{\omega_0 MS}{\delta_1 \delta_2 (1 + \beta^2)} \quad (2-158)$$

For the approximate estimate of signal lag in the amplifier of radio frequency it is possible to use an expression of time lag

$$t_s = - \left(\frac{d\varphi}{d\omega} \right)_{\omega=\omega_0} \quad (2-159)$$

where $\varphi = \varphi(\omega)$ - the expression of phase response.

Using expression (2-158), we obtain the shortened expression for

the phase response of the cascade/stage

$$\varphi = \operatorname{arctg} \frac{b_1 b_{20} (1 + \beta^2) - 4 \left(\frac{\Omega}{\omega_0} \right)^2}{2 (b_1 + b_{20}) \frac{\Omega}{\omega_0}}. \quad (2-160)$$

Then time lag tentatively can be considered according to the formula

$$t_0 = - \frac{d\varphi}{d\omega} \Big|_{\omega=0} = - \left(\frac{d\varphi}{d\Omega} \right) \Big|_{\Omega=0} = \frac{2 (b_1 + b_{20})}{\omega_0 (b_1 b_{20} + k_{cs}^2)}. \quad (2-161)$$

With change δ_{cs} time lag is constant with satisfaction of the condition

$$\frac{dt_0}{d\delta_{20}} = 0. \quad (2-162)$$

Differentiating formula (2-161) and equalizing with zero, we obtain the condition of the constancy of time lag with a change in value δ ,

$$k_{cs} = \delta_1. \quad (2-163)$$

Then

$$t_0 = \frac{2}{\omega_0 \delta_1} = \text{const.} \quad (2-164)$$

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With the shunting of the first duct/contour the analogous

condition of the stabilization of time lag

$$k_{\text{cs}} = \delta_1 \quad (2-165)$$

is obtained.

The reasonings conducted are accurate for the linear system. For the nonlinear system of expression (2-163), (2-165) can be taken as the approximate conditions of stabilization, but not the constancy of time lag.

With the identical ducts/contours, when $\delta_1 = \delta_2$, and critical coupling between the ducts/contours

$$k_{\text{cs. np}} = \sqrt{\frac{\delta_1^2 + \delta_2^2}{2}} = \delta_1 = k_{\text{cs. k.}} \quad (2-166)$$

Thus, condition (2-163) is a condition of the critical coupling, in which the passband is determined by the expression

$$\Pi_{\text{K}} = \Pi_{\text{np}} = \frac{1}{\sqrt{2}} (\delta_1 + \delta_2). \quad (2-167)$$

With increase δ_{max} increases and the condition of critical coupling is broken. Coupling coefficient is reduced and passband is changed on the sufficiently complicated dependence on δ_{20} . Let us find the further conditions for the selecting δ_{20} and circuit dampings δ_1 and δ_2 , on the basis of the guarantee of constancy of passband.

In the general case passband

$$\Pi = f_0 \sqrt{V[k_{cs}^2 + \delta_1 \delta_{20}]^2 + [k_{cs}^2 - 0.5(\delta_1^2 + \delta_{20}^2)]^2 + k_{cs}^2 - 0.5(\delta_1^2 + \delta_{20}^2)}. \quad (2-168)$$

This expression can be converted

$$\psi = \frac{\Pi}{f_0 \delta_1} \sqrt{V(\phi + q)^2 + (\phi - \frac{q^2 + 1}{2})^2 + \phi - \frac{q^2 + 1}{2}}, \quad (2-169)$$

where

$$\phi = \frac{k_{cs}^2}{\delta_1^2}; \quad q = \frac{\delta_{20}}{\delta_1}.$$

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It is necessary to find this value of parameter $\phi = \frac{k_{cs}^2}{\delta_1^2}$, at which passband would be changed least of all with change δ_{20} . This value

ϕ can be determined from the graphs in Fig. 37, 1,

in which relationships $\frac{\Pi}{f_0 \delta_1} = \psi(q)$ are plotted for different values of ϕ . The graph shows points corresponding to the values of q_{np} , at which the link is critical. The broken line connecting these points divides the graph into two domains in which one of the resonance curves has one peak, while in the other it has two.

From the graph it is evident that the most favorable value of parameter ϕ lies/rests within limits of 0.5-0.8. At these values ϕ an increase of the attenuation δ , 8 times changes passband not

more than 10%. To values $\phi = 0.6-0.8$ (point 1 and 2 in Fig. 37) it corresponds the value of the relations of the initial attenuations

$$q_n = \frac{\delta_2}{\delta_1} = 0.5 + 0.8.$$

Thus, so that with the shunting of secondary circuit the passband would be changed insignificantly, necessary to satisfy the condition

$$\delta_2 = q_n \delta_1 = (0.5 + 0.8) \delta_1. \quad (2-170)$$

or

$$Q_{10} = q_n Q_{20} = (0.5 + 0.8) Q_{20}.$$

i.e. the energy factor of the first duct/contour must compose 0.5-0.8 from the initial energy factor of the unshunted secondary circuit. Apparently, the most favorable relationship/ratio when $q_n = 0.63$.

So that resonance curve of amplifier would be flat/plane, it is necessary with the work of amplifier in the linear conditions the connection/communication between the ducts/contours to choose critical. Then initial passband is determined according to expression (2-167).

For determining the law of a change in conductivity σ_{res} let us rewrite expression (2-158) in the form

$$K_0 = \frac{\omega_0 S \sqrt{L_1 L_2} A}{\delta_2 + \delta_1 A^2}. \quad (2-171)$$

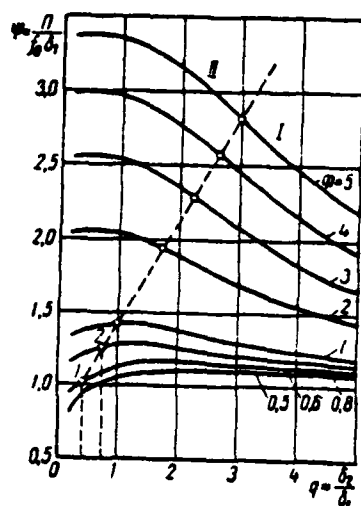


Fig. 37. The graph/diagrams of dependence $\psi(q)$ for the different values of the parameter f : I - region of single-humped resonance curve; II - region of two-humped resonance curve.

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In this case the initial factor of amplification of cascade/stage with the work in the linear conditions

$$K_u = \frac{\omega_0 S \sqrt{L_1 L_2} A}{b_1 (1 + q_u A^2)}. \quad (2-172)$$

On the basis of expression (2-19) we write/record the equality

$$K_0 = K_u \frac{1}{A}. \quad (2-173)$$

Taking into account that $g_{01} = q_u g_{01}$ and deciding relatively g_{u01} , we obtain

$$g_{u01} = g_{01} \left\{ m \left[\frac{2}{3} (A^3 + q_u) - A^3 \right] - q_u \right\}, \quad (2-174)$$

where $m=C_2/C_1$; $A = \sqrt{0.5(1+q^2)}$.

Being assigned by value q_m on the graph in Fig. 37 we determine value f for the critical coupling (points on the broken line). It is possible to be assigned by the value f and to determine q_m in the critical coupling. For example, with $\phi=0.7$ through the graph we find $q_m=0.6$.

From expression (2-174) it is possible to calculate dependence $g_{max}=-f(z)$, necessary for obtaining FAKh of any type.

§4. Some special features/peculiarities of FU on the transistors.

Special features/peculiarities of FU on the transistors are defined by the physical properties of transistor as amplifier instrument. In comparison with vacuum lamp it is possible to note the following fundamental special features/peculiarities of transistor:

transistor is substantially nonlinear amplifier instrument and works with comparatively small input and output voltages of the reinforced signal, which impedes the selection of the nonlinear element/cell, which forms FAKh of amplifier;

Input and output conductances of transistor considerably differ by value, in consequence of which for obtaining the sufficiently high transmission factor of cascade/stage the load must ensure agreement between the transistors, i.e., the load of amplifier stage must be matching device;

transistor is inertial amplifier instrument with the complex parameters, which depend on the frequency.

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Therefore in the selective amplifiers, in the first place, the resonance frequency of amplifier does not coincide with the resonance frequency of oscillatory system, and, in the second place, in the amplifier can be observed, besides fundamental resonance, further resonances at frequencies, which differ from fundamental signal frequency, which leads to the distortion of resonance curve of amplifier;

transistor has the large reverse/inverse transfer admittance Y_{11} , as a result of which in the selective amplifiers descends the coefficient of stable amplification and is distorted the form of

resonance curve of input and output circuits;

the dependence of the parameters of transistor, including of slope/transconductance of passage characteristic, input and output conductances (active and reactive/jet components) on the operational conditions of transistor (strength of current of emitter and voltage/stress on the collector/receptacle).

The enumerated special features/peculiarities of transistor cause the specific character of the circuit solution and construction of the transistors.

In the majority of the practical amplifier circuits of radio frequency the common-emitter connection (OE) is used, which has larger amplification than common-base circuit (OB), as a result of the larger input resistance. At comparatively low frequencies (for this type of transistor) the common-emitter connection considers fundamental. However, with the increase of frequency the difference in the values of the input resistances of common-emitter connections and by base is reduced.

At the sufficiently high (for this type of transistor) frequencies the amplifier with the general/common/total base can be almost also effective as common-emitter amplifier. Furthermore, in

the amplifier in OB internal feedback is exhibited more weakly, in consequence of which the amplifier works more stably. Therefore diagram with OB in many instances for the broadband selective amplifiers can prove to be preferable.

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Linearization of the passage characteristic of transistor.

The passage characteristic of transistor it is possible to linearize by linear effective resistance, including it:

a) in input circuit of cascade/stage consecutively/serially with the input electrode of transistor (Fig. 38a, b);

b) in the circuit of consecutive current feedback of emitter in the diagram with OE (Fig. 38c), on the base current in the diagram with OB (Fig. 38d);

c) in the circuit of parallel voltage feedback between the output and input terminals of transistor (Fig. 38e).

Each of the enumerated diagrams has their advantages and disadvantages.

As shown to work [8], the linearizing action of resistor/resistance R_b in $(\beta+1)$ is more than the linearizing action of resistor/resistance R_e . Upon the inclusion of the linearizing resistor/resistance into the circuit of base the temperature operational stability of transistor on the direct current descends, what is essential deficiency/lack.

The sufficiently high input resistance of cascade/stage at the high frequencies is the advantage of diagrams a and b. To deficiencies/lacks in the diagrams should be related: the decrease of transmission factor on the voltage/stress with the previous frequency properties of cascade/stage. Diagrams a and b are applied rarely, in essence at the high frequencies.

More frequently is applied the diagram, represented in Fig. 38c, advantage of which it is:

an increase in the input resistance at frequencies $f < f_\beta$

a partial improvement in the frequency properties of transistor, since according to work [22] the time constant of diagram with current feedback of the emitter

$$\tau_{02} = \frac{1 + \frac{R_b}{r_e}}{1 + R_b(S_e + \beta)} \tau < \tau.$$

Deficiencies/lacks in the diagram:

the decrease of the slope/transconductance of the passage characteristic of transistor and, consequently, also the transmission factor of cascade/stage to $1 + SR_0$ once;

at frequencies $f > f_0$ feedback becomes complex; therefore the linearizing action of feedback is reduced and input resistance is reduced.

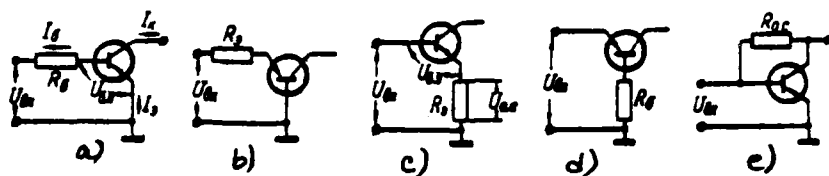


Fig. 38. Circuit solutions of the linearization of passage characteristic.

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The first deficiency/lack can be partially reduced, increasing the load resistance/resistor to the value, which ensures the previous frequency properties of amplifier, which the cascade/stage before the inclusion/connection of resistor/resistance R_0 . The scheme can be applied in pulse aperiodic and selective FU during the amplification of signal at frequencies $f < f_0$.

In work [8] it is shown that with the value of resistor/resistance $R_0 = 100$ ohms the high-frequency transistors of types P401-P403, P411, P417 and P418Zh can be counted as linear UP to input voltage $U_{\text{вх}} \approx 200$ mV, when $R_0 = 200$ ohm - to $U_{\text{вх}} \approx 0,4$ V, when $R_0 = 500$ ohm - to $U_{\text{вх}} \approx 0,8$ V.

Diagram in Fig. 38e, with voltage feedback considerably improves the frequency properties of cascade/stage. With carrying out of feedback on the diagram in Fig. 38e input and output conductances are changed, and this affects the transmission factor of the previous cascade/stage. In the cascade/stage in question it virtually does not

vary, since it is determined by the ratio of voltage/stress on the collector/receptacle to the voltage/stress on the base, that are internal for the closed loop of feedback. Therefore it is necessary to examine the pair of cascades/stages.

In a deep feedback the frequency characteristic of cascade/stage is distorted because the feedback loop forms the frequency-dependent voltage divider, which consists of effective resistance R_{∞} and complex input admittance \dot{Y} [22]. Therefore feedback is frequency-dependent. For obtaining the frequency-dependent feedback it is necessary to include impedance [22] selected by correspondingly.

MATCHING AND DECOUPLING OF AMPLIFIER STAGES ON THE TRANSISTORS.

Aperiodic FU. In the aperiodic transistorized amplifiers the cascades/stages between themselves usually do not coordinate. Therefore as basis applies diagram with OE, in which input and output conductances in the least measure are distinguished between themselves. For facilitating the adjustment multistage FU, which consists of the nonlinear cascades/stages, between the nonlinear cascades/stages it is expedient to include the untying emitter followers. In this case during the calculation of nonlinear cascade/stage by the effect of the subsequent nonlinear cascade/stage

it is possible to disregard.

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Practical diagrams of aperiodic FU, in particular logarithmic transistorized amplifiers, and their detailed research are given in work [8]. The LAKh in these amplifiers is obtained by the shunting of collector loads by nonlinear elements/cells or by nonlinear dividers. Accuracy of LAKh is $\pm 3\%$, slope/transconductance $\sigma = 3-3.5$ mv/dB. In all three diagrams for the realization of the successive work of nonlinear cascades/stages and obtaining LAKh of high accuracy between the nonlinear cascades/stages are connected the untying emitter followers.

Virtually instantaneous restoration/reduction of the sensitivity of amplifier after the break-down of large signals in entire dynamic range of LAKh and obtaining in the amplifier of the stable delay time of signal with a change of its value in entire range of LAKh of amplifier is the advantage of amplifier with 2nd type nonlinear dividers.

Resonance of FU. In resonance FU the single oscillatory circuits, tuned for one frequency, are applied. For decreasing the shunting effect of output and input admittances considerable in the

value the undercoupling of duct/contour with the transistors is applied. The basic versions of the circuit diagrams of duct/contour are shown in Fig. 39. On the diagrams the current generator $I = \dot{U}_{\text{ex}} Y_{21}$ and output conductance $Y_{22} = (1/R_{22}) + j\omega C_{22}$ are the equivalent of the previous transistor, by the equivalent of the subsequent transistor - input admittance $Y_{11} = (1/R_{11}) + j\omega C_{11}$. Conductivities Y_{21} , Y_{22} and Y_{11} correspond to the circuit diagrams of the transistor: OE, OB and OK.

The degree of the inclusion of duct/contour into output circuit of previous and into input circuit following of transistors is characterized by transformation ratios

$$m_1 = \frac{U_1}{U_n}; \quad m_2 = \frac{U_2}{U_n},$$

which are less than one.

Consequently, the greatest voltage of signal is created on the oscillatory circuit, i.e., voltage/stress U_n . This must be remembered during the selection of connection point into the load of the shunting nonlinear element/cell.

It is known that the required nonlinear element/cell to fit the easier, the greater voltage of signal U_{non} applied to the nonlinear element/cell.

In this case, connecting in series with the nonlinear element/cell supplementary resistors/resistances different in the value, it is possible to use a broad class of the semiconductor diodes, which have both the large and comparatively small slope/transconductance of the initial section of volt-ampere characteristic. Is most expedient the inclusion/connection of nonlinear element/cell shown by continuous duct/contour.

For the low-frequency diagrams it is possible to use any of four circuit diagrams of duct/contour (Fig. 39) independent of the method of the start of transistor. It is possible to apply the optimum coupling of duct/contour with the transistors and to obtain the maximum of amplification factor. The procedure of calculation of transformation ratios for the optimum coupling of duct/contour with transistors is given in work [67].

To calculate and design the low-frequency circuit and is comparatively easy. To much more complicatedly design amplifier at frequencies it is higher than 10 MHz, especially broadband.

Complexity is caused by the fact that the reactive components of input and output resistance of transistors are congruent in the value with the virtually realizable reactances of oscillatory circuits.

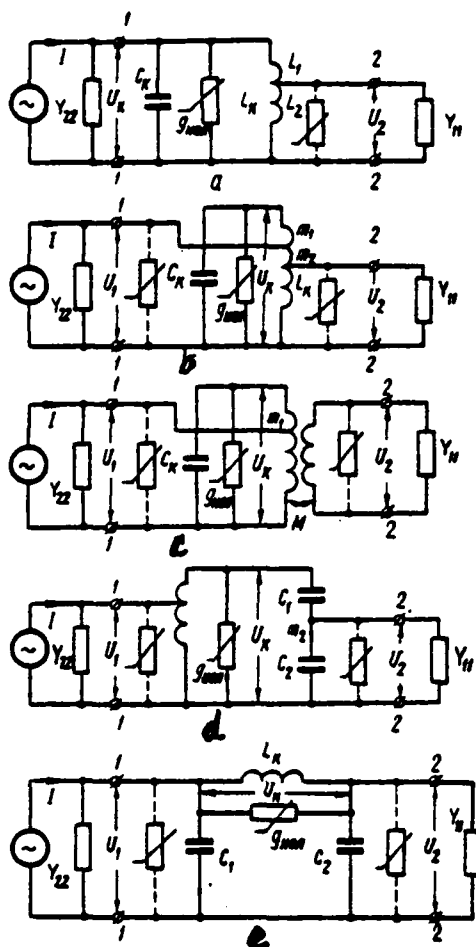


Fig. 39. Basic concepts of the inclusion/connection of oscillatory circuit and shunting nonlinear element/cell in single-tuned coupling: a) autotransformer; b) double autoinductive; c) transformer; d) capacitive; e) with series inductance.

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With the incomplete connection of transistors to the duct/contour

appear spurious resonances, whose frequencies do not coincide with the resonance frequency of fundamental duct/contour. In this case is formed the system, which during the appropriate selection of the points of connection can have flat/plane or double-humped resonance curve.

Spurious resonances can be used in the general case for the correction of the frequency characteristic of amplifier for obtaining its flat/plane or double-humped form. However, this method of the correction of frequency characteristic, as shown in work [8], is not suitable with the shunting of oscillatory circuit by nonlinear element/cell.

Spurious resonances can be reduced, selecting the appropriate points of the connection of transistors to the oscillatory circuit and applying the capacitive voltage-divider in the diagram with OE and the autoinductive inclusion of duct/contour in the diagram with OB.

The distortion of the resonance curve selective system over the dynamic range can be reduced, connecting two nonlinear elements/cells to in parallel to the basis LC, and parasitic $L_0 C_{0x}$ oscillatory circuits. The selection of such nonlinear elements/cells is sufficiently complex, since the nonlinear elements/cells must be

different. However, when it is necessary to obtain wide passband in the amplifier on the single ducts/contours, this method of the distortion elimination of resonance curve is completely acceptable.

Let us find the required law of a change in the conductivity of nonlinear element/cell. During the analysis of high-frequency amplifier in general form, taking into account the effect of self-feedback, the bulky formulas, unsuitable for the engineering calculations, are obtained. Therefore for simplification in the linings/calculations by analogy with amplifiers on vacuum lamps during the analysis we will assume that the self-feedback in the transistor is neutralized, i.e., conductivity $Y_{12}=0$.

Independent of the method of the inclusion of duct/contour into output and input circuits of transistors calculation formulas are one and the same. Therefore let us consider fundamental calculated relationships/ratios based on the example of diagram with the double autoinductive inclusion of the duct/contour (Fig. 39b), in parallel to which is connected nonlinear element/cell with the transformation ratio $m=1$.

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The complex transmission factor of cascade/stage on the

voltage/stress

$$K = \dot{S}\dot{Z}_0 = \frac{S_0 R_0' m_1 m_2}{(1+r)(1+p)}, \quad (2-175)$$

where S_0 - low-frequency value of the slope/transconductance of direct drive; $p = \frac{y}{d_0} = yQ_0$ - generalized detuning; $d_0 = \frac{p}{R_0}$ - complete attenuation; $y = (\omega/\omega_0) - (\omega_0/\omega) \approx \Delta\omega/\omega_0$ - detuning; $\Delta\omega \approx \omega - \omega_0$; $\omega_0 = 1/(\sqrt{LC_0})$; $r \approx \omega\tau = \frac{\omega}{\omega_{rp}}$; ω_{rp} - the cut-off frequency of the transistor, connected on the appropriate diagram; $R_0 = \frac{1}{g_0}$ - equivalent resonance resistor/resistance of the shunted duct/contour

$$\left. \begin{aligned} R_0 &= \frac{R_0 R_{\text{нел. c}}}{R_0 + R_{\text{нел. c}}}; \\ \frac{1}{R_0} &= \frac{1}{R_{0c}} + \frac{m_1^2}{R_{22}} + \frac{m_2^2}{R_{11}}; \end{aligned} \right\} \quad (2-176)$$

R_{0c} - the equivalent resonance resistor/resistance of duct/contour.

The total capacitance of the duct/contour

$$C_0 = C_k + m_1^2 C_{22} + m_2^2 C_{11} + C_w + C_{\text{нел. c}}$$

where C_k - the self-capacitance of duct/contour; C_w - wiring capacitance; $C_{\text{нел. c}}$ - capacity/capacitance of nonlinear element/cell.

Modulus/module of transmission factor

$$K = \frac{S_0 R_0' m_1 m_2}{\sqrt{(1+r^2)(1+p^2)}}. \quad (2-177)$$

Frequency characteristic

$$A = \frac{K}{K_0} = \frac{1}{\sqrt{(1+r^2)(1+p^2)}}. \quad (2-178)$$

The frequency and phase responses of transistorized amplifier in the general case are asymmetric relative to the resonance frequency

of oscillatory circuit ω_0 . The degree of asymmetry depends on the quality of duct/contour and relationship/ratio of frequencies $\frac{\omega_0}{\omega_{rp}} = \xi$.

Equalizing the imaginary part of expression (2-175) zero, we find the frequency, at which the phase of system is equal to zero

$$\omega_{\varphi=0} = \frac{\omega_0}{\sqrt{1 + \frac{\xi}{Q}}}. \quad (2-179)$$

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The value of frequency ω_m , at which the transmission factor is maximum, we find from equation $\frac{dK}{d\omega} = 0$

$$\omega_m = \sqrt{S - \frac{b}{3a}}. \quad (2-180)$$

where

$$s = u + v; \quad u = \sqrt{-m + \sqrt{m^2 + n^2}}; \quad v = \sqrt{m - \sqrt{m^2 + n^2}};$$

$$m = \frac{b^2}{27a^2} + \frac{d}{2a}; \quad n = -\frac{b^2}{9a^2}; \quad a = 3\xi^2; \quad b = 2(\xi^2 + Q^2 - 2Q^2\xi^2); \quad d = 2Q^2\omega_0^4.$$

Formula (2-179) is precise, but on it is impossible to judge graphically the change in frequency ω_m with a change in quality Q .

For the approximate determination of frequency ω_m with $\varphi=0$ it is possible to use following approximation formula [51]:

$$\omega_m = \omega_0 \sqrt{1 - \frac{1}{Q^2}}. \quad (2-181)$$

Initial transmission factor of cascade/stage with the work in the linear conditions

$$K_n = \frac{S_0 R_0 m_1 m_2}{\sqrt{(1 + r_m^2)(1 + p_{n,m}^2)}}, \quad (2-182)$$

where

$$r_m = \omega_m \tau_i; \quad p_{n,m} = Q_n \left(\frac{\omega_m}{\omega_0} - \frac{\omega_0}{\omega_m} \right) = Q_n y_m.$$

With the work of cascade/stage in the functional and quasi-linear modes/conditions transmission factor

$$K = \frac{S_0 R_0 m_1 m_2}{\sqrt{(1 + r_m^2)(1 + p_m^2)}}, \quad (2-183)$$

where $p_m = Q_0 y_m = \frac{y_m}{d_0}$.

Taking into account condition (2-19), we find expression for the conductivity of the nonlinear element/cell

$$g_{\text{нел}} = g'_0 \left(\sqrt{(1 + p_{\text{н.н}}^2) \left(\frac{x}{z} \right)^2 - \frac{y^2}{p^2 (g'_0)^2}} - 1 \right). \quad (2-184)$$

For the narrow-band amplifiers, for which is fulfilled inequality $Q \gg 1$, and broadband when $\omega_{rp} > \omega_0$, value $y_m \rightarrow 0$ and expression (2-184) is converted into expression (2-115).

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Thus, conductivity and the volt-ampere characteristic of nonlinear element/cell can be calculated from expressions (2-115) and (2-117), i.e.

$$g_{\text{нел}} = g'_0 \left(\frac{x}{z} - 1 \right); \quad I_{\text{нел}} = U_{\text{нел}} g_{\text{нел}},$$

where $g'_0 = g_0$; $U_{\text{нел}} = \frac{U_{\text{вмх}}}{m^2} = U_{\text{вмх}} \frac{z}{m_2}$ - with the complete connection of nonlinear element/cell to the duct/contour; $g'_0 = g_0 / m^2$; $U_{\text{нел}} = U_{\text{вмх. нз}}$ - with the incomplete connection of nonlinear element/cell from exit

side of transistor; $g'_0 = g_0/m^2$; $U_{\text{non}} = U_{\text{max}} \cdot z^2$ - with the incomplete connection of nonlinear element/cell from entry side of the following transistor (z - standardized/normalized output voltage/stress).

During the fluent analysis of expressions (2-176) and (2-178) it seems that the asymmetry of the resonance and phase responses of nonlinear cascade/stage with an increase in the signal must grow, since the energy factor of duct/contour is reduced. However, this not thus. The asymmetry of frequency and phase responses, and also passband in the nonlinear cascade/stage are less in comparison with the linear, that has $g_{0, \text{non}}$ equal to $g_{0, \text{non}}$ of nonlinear cascade/stage on this signal level and at frequency ω_m . This phenomenon can be explained by the fact that with any change in the voltage of signal on the oscillatory circuit in proportion to detuning the action of nonlinear element/cell is such, that it reduces this change as a result of a change in the conductivity of nonlinear element/cell g_{non} , and consequently, g_0 . To calculate the resonance characteristic of nonlinear cascade/stage is very difficult, since in this case it is necessary to solve transcendental problem.

Let us pause at the selection of transformation ratios m_1 and m_2 with the work of cascade/stage in the linear conditions.

In work [22] it is shown that with the assigned passband the

maximum amplification is obtained with the equality of those introduced into the duct/contour of the resistors/resistances

$$R'_{11} = R'_{22}, \text{ or } \frac{R_{11}}{m_1^2} = \frac{R_{22}}{m_2^2}. \quad (2-185)$$

When the maximum factor of amplification of cascade/stage K_{max} exceeds the value of the actually realizable dynamic range of LAKh in one cascade/stage, the mode/conditions the agreements reject. In this case the complete inclusion/connection of duct/contour from exit side ($m_1=1$) is applied, and value of m_2 is chosen from the condition of the permissible shunting of duct/contour with the input resistance R_{11} of the following cascade/stage.

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At the high values of input capacitance C_{11} is sometimes more expedient value m_2 to take small, and the required passband to obtain, connecting in parallel to duct/contour back-out resistor R_m . In this case the effect of the instability of capacity/capacitance C_{11} on the passband sharply descends. The method of expanding the passband examined it is expedient to apply in the amplifier with the ARU due to a change in the mode/conditions of the work of the transistor, whose input capacitance considerably is changed in the alignment procedure of amplification.

In work [2] it is shown that the efficiency

$$\eta = \Pi_n K_n$$

of broadband cascade/stage is greatest at the completely specific values of coefficient of m_n and its shunting resistor/resistance R_{sn} . The most fully possible cases of the selection of coefficients m_n and R_{sn} are examined in work [67].

DECREASE OF THE REVERSE/INVERSE TRANSFER ADMITTANCE $Y_{1,1}$.

Parasitic feedback can be considerably weakened/attenuated, using transistors with a small reverse/inverse transfer admittance (transistors with four electrodes p-n-i-p); neutralizing the self-feedback of transistor by means of external feedback; applying the composite/compound cascode circuit diagrams of transistors.

Applying special transistors with a small reverse/inverse transfer admittance and composite/compound circuit diagrams of transistors, it is possible to obtain stable amplification in the broadband. Therefore it is expedient to apply the first two methods of decreasing the parasitic feedback in the wideband amplifiers.

During the neutralization of self-feedback the stability of

amplification in a comparatively narrow frequency band is raised. Because of this the neutralization is applied only in the narrow-band amplifiers.

Questions of the neutralization of feedback in the high-frequency amplifiers (UVCh) on the transistors are in sufficient detail examined in works [42, 47, 52, 68].

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The neutralization of self-feedback in the transistorized amplifiers is realized, connecting to the amplifier the special linear and passive neutralizing quadrupoles. The frequency and phase responses of the neutralizing quadrupole are chosen so that the external feedback would neutralize internal connection in the frequency band.

The diagrams of neutralization can be reduced to four circuits; series (neutralization of the type Z); parallel (neutralization of the type Y); series-parallel (neutralization of the type h); parallel-series (neutralization of the type g) or to their combinations.

In the transistorized amplifiers there is a specific character

in the use/application of diagrams of neutralization. If dynamic range is reached with the help of the gain control by a change in the voltage/stress on collector/receptacle U_k or the emitter current of transistor I_e , modulus/module and phase of the coefficient of the feedback of transistor are changed. Because of this the calculation and the selection of the elements/cells of the neutralizing quadrupole, which efficiently functions in the broad dynamic band, hinders. The elements/cells of the neutralizing quadrupole must be calculated, on the basis of the average/mean value of modulus/module and phase of the coefficient of feedback in the dynamic range.

For comparatively narrow-band amplifiers with ShDD, as shown in work [8], a diagram of the type Y is the most acceptable diagram of neutralization.

In work [68] it is shown that with the increase of frequency the coefficient of the feedback of transistor at first is raised, and then descends and at the sufficiently high frequency for this type of transistor amplifier it works stably. During the appropriate selection of the connection/communication of transistor with the input and output circuits (with the signal generator and the load) the need for neutralization is eliminated. Therefore the creation of the serial wideband amplifiers of the oscillations/vibrations of radio frequency with the ShDD is facilitated independent of the

method of gain control.

USE OF CASCODE CIRCUITS IN FUNCTIONAL AMPLIFIERS.

Is called the cascode diagram which consists of two amplifier instruments UP₁ and UP₂ (Fig. 40), each of which can be included/connected on one of three basic concepts: with the general/common/total ones by base (OB), by emitter (OE) and by collector/receptacle (OKp) for the transistors or with the general/common/total ones by grid (OS), by cathode (OKд) and by the anode (OA) for vacuum lamps.

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Nine combinations of the circuit diagrams of the first and second amplifier of instruments are theoretically possible. For the transistors:

OB — OB	OB — OЭ	OB — OKp
OЭ — OB	OЭ — OЭ	OЭ — OKp
OKp — OB	OKp — OЭ	OKp — TKp

For vacuum lamps:

OC — OC	OC — OKд	OC — OA
OKд — OC	OKд — OKд	OKд — OA
OA — OC	OA — OKд	OA — OA

In the cascode diagrams the load between the amplifier instruments, as a rule, is absent and there is a conductive coupling.

Because of this the combinations of amplifier instruments indicated can be considered as common UP with the equivalent parameters, determined by characteristics and circuit diagrams of the entering it transistors.

Cascode diagrams great are applied in the transistor amplifiers.

To the analysis of the linear amplifiers, made during the combinations of transistors, it is devoted the series/row of works [6, 8, 51, 64, 91].

Let us consider the diagrams of power feed of transistors and cascode diagrams from the point of view of the expansion of the dynamic range of amplifier.

Depending on the method of the start of transistors by the direct current and grid priming the diagrams with parallel, series and series-parallel feed are distinguished.

With parallel feed (Fig. 41a) it is possible by more bending to choose the mode/conditions of transistor on the direct current and the method of its adjustment. With the series feed (Fig. 41b) a number of parts is decreased and current drain is reduced from the power supply in comparison with the parallel version of feed.

The need of applying the power supply with the increased voltage/stress is a deficiency/lack in the series feed. However, the deficiency/lack indicated in certain cases can be realized as advantage. In particular, with those increased supply voltage and resistance/resistor of the load of cascade/stage, it is possible to obtain the large signal amplitude, the selection of the nonlinear element/cell, which shunts oscillatory circuit, is facilitated by this.

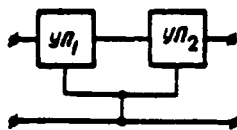


Fig. 40. Block diagram of the cascade start of two amplifier instruments.

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The more the series-connected transistors, the more supply voltage is required, the more possible it is to obtain the voltage of signal on the duct/contour. With the series-connected transistors the maximum amplitude of the voltage of signal on the oscillatory circuit can be obtained in n of times more than in the cascade/stage with one transistor.

It is natural that this can be obtained as a result of an increase of the load resistance/resistor in n of times, which will lead to the contraction of the passband of tuned amplifier approximately/exemplarily in n of times. Consequently, the method of increasing the amplitude of output voltage/stress examined can be realized in the narrow-band resonance or aperiodic pulse amplifiers.

If in the amplifier transistors are included/connected on the

diagram OE- OB-OB...-OB, then frequency properties of the equivalent UP will be determined by the frequency properties of the first transistor, connected on the diagram OE. Therefore in the aperiodic amplifier with an increase in the load resistance/resistor the frequency properties of cascade/stage in effect will not deteriorate until the inequality

$$C_0 R_0 < \tau_n$$

is fulfilled where C_0 and R_0 - capacity/capacitance and the total resistance of load.

With high-amplitude onset of the voltage of signal on the load the breakdown between collector/receptacle and base of latter/last transistor can occur. For eliminating this phenomenon it is necessary to carry out feedback on the alternating voltage between collector/receptacle and base of latter/last transistor (Fig. 41).

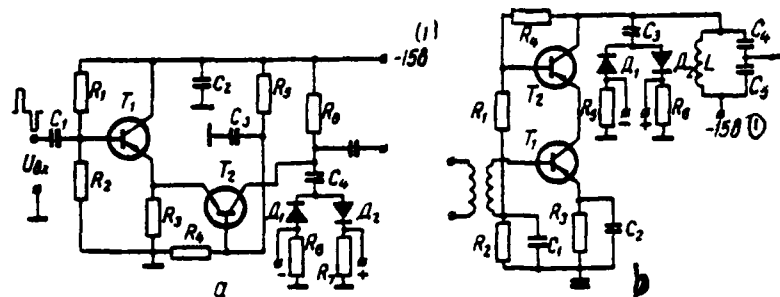


Fig. 41. Cascode circuit diagrams of transistors with different feed on the direct current: a) parallel; b) consecutive.

Key: (1). V.

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In this case the potential between collector/receptacle and base of transistor remains constant/invariable, and transistor works reliably.

During the use of transistors of two types the p-n-p and n-p-n efficiency of compound configuration can be raised, applying the more economical mode of feeding of transistors. To increase the level of the output voltage of the signal is possible in the push-pull transformerless diagram, which consists of the transistors of types p-n-p and n-p-n. In this diagram the energy consumption from the power supply in the absence of signal can be reduced to a minimum

[22].

S 5. Functional amplifiers with the nonlinear feedback.

To regulate the transmission factor of cascade/stage with an increase in the signal is possible with the help of the active nonlinear negative feedback on alternating current or voltage/stress.

FU WITH NONLINEAR FEEDBACK ON THE CURRENT.

The simplified circuits of amplifier stages with the nonlinear negative feedback (NOOS) on the alternating current component of common electrode UP are depicted in Fig. 42. The NOOS is realized by inclusion/connection in parallel to the resistor/resistance of feedback R of the nonlinear element/cell, whose resistor/resistance (conductivity) is changed with an increase on it in the voltage of signal $U_{\text{non. o'}}$

On the constant component nonlinear element/cell it is isolated from the cathode (emitter) by isolating capacitor C_p .

At the high frequencies ~ 10 MHz in the vacuum-tube amplifiers considerable by-passing of stray capacitance in the cathode circuit of cascade/stage appears. Because of this the active nonlinear

element/cell, shunted by stray capacitance, ceases to affect the transmission factor of cascade/stage, since at the high frequencies there is no negative feedback.

For eliminating this phenomenon into the cathode circuit of cascade/stage it is necessary to include the oscillating duct/contour, tuned to a frequency of the reinforced vibrations.

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During the calculation of the inductance of cathode oscillatory circuit one should consider that the value of stray capacitance between cathode and filament for different types of tubes varies from 3 to 10 pF, and capacitance value of mounting in the cathode circuit can reach 5-10 pF.

In the transistor amplifiers at frequencies $f > f_0$ (for the diagrams with OE) are exhibited the inertial properties of transistor, feedback is complex and the effect of adjustment sharply descends. For eliminating the phenomenon indicated in the feedback loop is connected corrective capacity/capacitance C_n (Fig. 42b), which together with the internal inductance of emitter conclusion/output forms consecutive oscillatory circuit. Further isolating capacitor in this case is not required.

The internal inductance of the emitter conclusion/output

$$L_{\text{em}} = L_{\text{u}} + L_{\text{r}}, \quad (2-186)$$

where L_{u} — structural/design inductance of emitter circuit; L_{r} — internal diffusion inductance of emitter circuit, determined by the inertial properties of the transistor

$$L_{\text{r}} = \frac{r_0}{2\pi f_c}. \quad (2-187)$$

If we take for the transistor of the type P403 $r_0 = 50$ ohm, $f_c = 200$ MHz, then $L_{\text{r}} \approx 0.04$ μH .

Measured structural/design inductance for high-frequency drift diffused-bases transistor of types P403, P410

$$L_{\text{u}} \approx 0.02 + 0.03 \mu\text{H}.$$

From expression (2-187) it is evident that with increase inductance L_{r} is reduced. Therefore at very low value L_{r} in the emitter circuit it is possible to include further inductance.

Nonlinear element/cell is connected with the series circuit.

In the presence of oscillatory circuit in the feedback loop of oscillatory circuit in the load circuit of selective cascade/stage it

can and cannot be.

In the electron-tube selective amplifiers, as a rule, in the feedback loop applies parallel circuit, in the transistor amplifiers - consecutive.

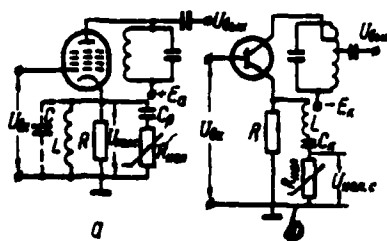


Fig. 42. Simplified circuits of FU with nonlinear current feedback:
a) on the tube; b) on the transistor.

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Diagram with the active resistor/resistance or the parallel oscillatory circuit in circuit of feedback. The complex transmission factor of the cascade/stage

$$\dot{K} = \frac{SZ_{\text{нагр}}}{1 + SZ_{\text{о.с}}}, \quad (2-188)$$

where $Z_{\text{нагр}}$ - impedance of load; $Z_{\text{о.с}}$ - impedance of the feedback

$$Z_{\text{о.с}} = \frac{R_{\text{о.с}}}{1 + j\rho_{\text{о.с}}}; \quad (2-189)$$

$g_{\text{о.с}} = \frac{1}{R_{\text{о.с}}} = \frac{1}{R} + \frac{1}{R_{\text{н}}} + \frac{1}{R_{\text{нагр}}}$ - complete conductance in the feedback loop; $R_{\text{н}}$ - resonance resistor/resistance of duct/contour;

$$d_{\text{о.с}} = \frac{\rho_{\text{о.с}}}{R_{\text{о.с}}}; \quad \rho_{\text{о.с}} = \sqrt{\frac{L_{\text{о.с}}}{C_{\text{о.с}}}} = \omega L_{\text{о.с}} = \frac{1}{\omega C_{\text{о.с}}}; \quad \rho_{\text{о.с}} = \frac{R}{d_{\text{о.с}}}.$$

If duct/contour in the feedback loop is absent, then $R_{\text{н}} = \infty$ and

$$g_{\text{о.с}} = \frac{1}{R} + \frac{1}{R_{\text{нагр}}}.$$

Modulus/module of the transmission factor of the cascade/stage

$$K = \frac{SR_{\text{о}}\Phi(R_{\text{нагр}})}{\sqrt{1 + \rho^2} \left(1 + \frac{SR_{\text{о.с}}}{\sqrt{1 + \rho_{\text{о.с}}^2}} \right)}. \quad (2-190)$$

In the presence of resonance $y=0$ transmission factor

$$K_0 = \frac{S R_0 \Phi(R)}{1 + S R_{0,0}} \quad (2-191)$$

Then the equation of resonance curve

$$A = \frac{K}{K_0} = \frac{1 + S R_{0,0}}{\sqrt{1 + p^2} \left(1 + \frac{S R_{0,0}}{\sqrt{1 + p_{0,0}^2}} \right)} \quad (2-192)$$

In the purely active feedback the phase response of amplifier is described by the known to expressions [59].

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With the work of cascade/stage in the linear conditions transmission factor

$$K_n = \frac{S_n R_0}{1 + S_n R_{0,0,n}} \quad (2-193)$$

where

$$g_{0,0,n} = \frac{1}{R_{0,0,n}} = \frac{1}{R} + \frac{1}{R_{\text{вс.н}}} = g_R + g_{\text{вс.н}} \quad (2-194)$$

since we consider that $R_n \gg R$.

Whence, at given values K_n , S_n and R

$$R_{\text{вс.н}} = \frac{(1 - \gamma) R}{\frac{R}{R_{\text{нар.н}}} - (1 - \gamma)} \quad (2-195)$$

where $\gamma = \frac{K_n}{S_n R_{\text{нар.н}}}$.

To determine the analytical expression of the law of a change in nonlinear conductivity $g_{\text{non. o}} = f(U_{\text{non. o}})$ with the work of cascade/stage in the nonlinear (functional and quasi-linear) mode/conditions is difficult, since voltage/stress $U_{\text{non. o}}$ depends on the value of the nonlinear resistance

$$U_{\text{non. o}} = U_{\text{ex. ex}} \frac{SR_{\text{o. o}}}{1 + SR_{\text{o. o}}} \quad (2-196)$$

In this case is obtained the complicated transcendental equation, which can be solved graphically for each specific case.

To much more simply separately find analytical expressions for dependences $g_{\text{non. o}} = f(U_{\text{ex}})$ and $U_{\text{non. o}} = \varphi(U_{\text{ex}})$, and from them for each specific case to calculate dependence $g_{\text{non. o}} = f(U_{\text{non. o}})$ or dynamic volt-ampere characteristic of nonlinear element/cell $I_{\text{non. o}} = f(U_{\text{non. o}})$.

Taking into account equations (2-191), (2-193) and (2-18), we find expression for the common conductivity of the feedback

$$g_{\text{o. o}} = \frac{1}{R_{\text{o. o}}} = \frac{S_{\text{u}} \Phi(S)}{\Phi(S) \Phi(R) \psi(x) (1 + S_{\text{u}} R_{\text{o. o. u}}) - 1}, \quad (2-197)$$

where $\psi(x) = x/z$, expressed through x .

Taking into account equation (2-194), we find

$$g_{\text{non}} = g_{\text{o. o}} - g_{\text{R}} \quad (2-198)$$

Introducing the standardized/normalized voltage of signal on the nonlinear element/cell, we obtain

$$x_{\text{non}} = \frac{U_{\text{non.o}}}{U_{\text{ex. n}}} = x \frac{SR_{\text{o.o}}}{1 + SR_{\text{o.o}}}. \quad (2-199)$$

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Taking into account formula (2-197), we obtain

$$x_{\text{non}} = \frac{x\phi(S)\phi(R)(1 + S_n R_{\text{o.o.n}}) - 1}{\phi(S)\phi(R)(1 + S_n R_{\text{o.o.n}})}. \quad (2-200)$$

In this case

$$U_{\text{non.o}} = U_{\text{ex. n}} x_{\text{non}} \quad (2-201)$$

Signal current, flowing through the nonlinear element/cell,

$$I_{\text{non.o}} = U_{\text{non.o}} g_{\text{non}} = U_{\text{ex. n}} \frac{(S - g_R [\psi(x)\phi(S)\phi(R)(1 + S_n R_{\text{o.o.n}}) - 1])}{\phi(S)\phi(R)(1 + S_n R_{\text{o.o.n}})}. \quad (2-202)$$

The dynamic volt-ampere characteristic of nonlinear element/cell can be constructed by the method of joint solution of equations (2-101) and (2-202).

Transmission factor of transistorized amplifier, connected on the diagram with OE, with negative current feedback accordingly [22]

$$K = \frac{S_{\text{os}}}{Y_{\text{narf}}} = \frac{S_{\text{narf}}}{1 + Z_{\text{o.o}}(S + Y)}. \quad (2-203)$$

With effective resistance $R_{\text{o.o}}$

$$K = \frac{S_0 R_0}{\sqrt{1 + (\omega_0 \tau_{\text{os}})^2} (1 + R_{\text{o.o}} S'_0)}, \quad (2-204)$$

where $S'_0 = S_0 + g$; $\tau_{\text{os}} = \tau \frac{1 + \frac{R_{\text{o.o}}}{r_0}}{1 + R_{\text{o.o}}(S_0 + g)}$; $r_{\text{o.o}} = r_0 + R_{\text{o.o}}$ - parameters of transistor at the presence of feedback. Usually $S_0 \gg g$ it is possible for the

approximate calculations to take $S'_{0.0} = S_{0.0}$ with resonance $p=1$.

Expression (2-197) correctly also for the transistor amplifier during the replacement in it S on $S'_{0.0}$. Frequency factor $\sqrt{1 + (\omega_0 \tau_{0.0})^2}$ in the denominator of expression (2-204), which considers the decrease of mutual conductance of transistor with an increase in the frequency, is decreased, since value g_{max} is found from the ratio of transmission factors $\frac{K}{K_n}$, which does not depend on frequency.

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Frequency factor must be calculated during calculation x_{max} on the basis of formula (2-199), substituting in formula $S = \frac{S_{0.0} \Phi(S)}{\sqrt{1 + (\omega \tau)^2}}$.

Diagram with the series circuit (Fig. 42b). The modulus/module of the transmission factor of the electron-tube resonance cascade/stage

$$K = \frac{SR_0}{\sqrt{1 + p^2(1 + SZ_{0.0})}}, \quad (2-205)$$

where $Z_{0.0} = \frac{RZ'_n}{R + Z'_n}$; $Z'_n = r' \sqrt{1 + p_{0.0}^2}$ - worsened/impaired resistor/resistance of series circuit; $r' = R_{loss} + r$; r - resistor/resistance of the losses of duct/contour; $p_{0.0} = yQ'_0$; $Q'_0 = \frac{p}{r}$ - the worsened/impaired quality of duct/contour in the feedback loop.

In the presence of resonance $y=0$

$$K = \frac{SR_0}{1 + SR_{0.0}}, \quad (2-206)$$

where $R_{0.0} = \frac{Rr}{R+r}$.

For obtaining the FAX conductivity $g_{0.0}$ must be changed according to the law (2-198). Then the resistor/resistance of the nonlinear element/cell

$$R_{\text{нел}} = \frac{1}{g_{\text{нел}}} = \frac{R_{0.0}(R+r) - Rr}{R - R_{0.0}}. \quad (2-207)$$

Standardized/normalized voltage/stress on the nonlinear element/cell

$$x_{\text{нел}} = x \frac{SR_{0.0}}{(1 + SR_{0.0})} \cdot \frac{R_{\text{нел}}}{R_{\text{нел}} + r}. \quad (2-208)$$

With $r \rightarrow 0$ expression (2-207) and (2-208) they are converted into expressions (2-198) and (2-199).

In the case of transistor amplifier at the series resonance frequency in the feedback loop expressions (2-207) and (2-208) are also valid. However, in this case in expression (2-208) it is necessary to substitute $S = S_0 / (\sqrt{1 + (\omega\tau)^2})$.

For the vacuum-tube amplifiers in many instances it is possible to count the load of the cascade/stage of constant and function $\Phi(R) = 1$. For satisfaction of condition $\Phi(R) = 1$ in the transistor

amplifiers it is necessary between the nonlinear cascades/stages to include the untying emitter followers.

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From expressions (2-197) and (2-207) it is evident that the law of a change of resisting the nonlinear element/cell and, consequently, also the form of its dynamic characteristic depend on slope/transconductance of UP, value of the resistor/resistance of feedback R and value of the voltage/stress of triggering/opening (closing) on the nonlinear of elements/cells, which determines initial value $R_{\text{нел. н.}}$ and, consequently, $R_{\text{о.о.н.}}$.

During the calculation of the dynamic volt-ampere characteristic of nonlinear element/cell and dependence $g_{\text{нел. о}} = f(U_{\text{нел. о}})$ it is expedient to construct functional dependences $g_{\text{нел. о}} = f(x)$ and $x_{\text{нел}} = \varphi(x)$, and on them - curves $g_{\text{нел. о}} = \eta(x_{\text{нел}})$ or $R_{\text{нел. о}} = \eta(x_{\text{нел}})$, which can be used for the construction of the volt-ampere characteristics of nonlinear element/cell or dependences $g_{\text{нел. о}} = f(U_{\text{нел. о}})$ with different values $U_{\text{нел. н.}}$. This construction is made in the work of the author [7, 8].

During the calculation of conductivity $g_{\text{нел. о}}$ to directly take into account a change in the slope/transconductance of UP in the

dynamic range is impossible, since conductivity $g_{\text{non. o.}}$, voltage/stress $U_{\text{non. o.}}$, voltage on control electrode of UP and mutual conductance of UP of mutually each other cause.

If passage characteristic of UP has considerable linear section, also, as a whole the nonlinearity of characteristic of UP small, then, by taking into account linearizing action of feedback, dependence $g_{\text{non. o.}} = f(U_{\text{non. o.}})$ it is possible to calculate first for $S = S_{\text{a}} = \text{const}$, and then to make more precise calculation with the large voltages of signal.

If the nonlinearity of passage characteristic is pronounced, dependence $g_{\text{non. o.}} = f(U_{\text{non. o.}})$ must be calculated by the method of iterations.

From expressions (2-197), (2-198) and (2-207) it is evident that for the realization of FAX with $b = \text{var}(\downarrow)$ is required the nonlinear element/cell with 2nd type characteristics, and with $b = \text{var}(\uparrow)$ - the 1st type.

A change in the input parameters in the dynamic range is one of the deficiencies/lacks FU with consecutive nonlinear current feedback. In particular, in vacuum-tube amplifiers the input capacitance of cascade/stage over the dynamic range is changed,

approximately/exemplarily, according to the following law:

$$C_{\text{ex. a. o.}} = \frac{C_{\text{ex}}}{1 + SR_{\text{a. o.}}}, \quad (2-209)$$

where C_{ex} — the input capacitance of amplifier stage without the feedback. This produces change in the total capacitance of the load of the previous cascade/stage.

In the transistor amplifiers the law of change $C_{\text{ex. a. o.}}$ is more complicated.

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In the broadband and selective amplifiers, when the capacity/capacitance of oscillatory circuit is small, with a change in capacity/capacitance C_{ex} it can considerably be changed the energy factor of duct/contour, frequency and phase responses. However, with the appropriate selection of the circuit elements of a change in the energy factor of duct/contour Q it is possible to reduce to a minimum.

The schematics of unipolar and bipolar logarithmic amplifiers with nonlinear current feedback on the tubes and the transistors are given in works [7] (Figs 63, 68), [8] (Figs 54, 58).

On the basis of the given theoretical and experimental studies

of logarithmic video amplifier with nonlinear current feedback of cathode (emitter) it is possible to do the following conclusions:

1) the range of LAKh of one cascade/stage it is possible to obtain to 28-30 dB;

2) if as the nonlinear elements/cells are used germanium diodes D2 or D9, then the greatest dynamic range of LAKh of cascade/stage is obtained during the voltages/stresses of triggering/opening $U_0=0.1\div0.15$ V and the resistors/resistances of feedback $R=3-6$ kilohm.

The LAKh of electron-tube cascade/stage begins with input voltage $U_{ex. \pi}=20\div30$ mV, transistor - 3-6 mV;

3) the slope/transconductance of LAKh of amplifier and its dynamic range it is easy to change, changing voltages/stresses U_0 on the nonlinear elements/cells;

4) in the multistage electron-tube and in the transistor amplifiers with the untying emitter followers it is possible to carry out a strictly successive work of nonlinear cascades/stages and as a result to obtain LAKh in the range 80-90 dB with the relative accuracy in voltage/stress $\delta U=3\%$ and in slope/transconductance $\delta \sigma=15-25\%$;

5) vacuum-tube amplifier is not overloaded, it does not lose sensitivity and reinforces signals according to the assigned law to 10-15 V, transistor amplifier - to units of volt;

6) the divergences of LAKh of a n-cascade amplifier, caused by aging and exchanging transistors, it is possible to eliminate the change in the bias voltages on the nonlinear elements/cells in the cascades/stages, in which were substituted the transistors or tubes;

7) amplifier stages are included by a deep direct-current feedback; therefore diagram has sufficiently stable parameters. Instability of LAKh with a change in the temperature in essence is determined by the instability of the parameters of nonlinear elements/cells. For increasing the temperature stability LAKh it is necessary as the nonlinear elements/cells to use silicon semiconductor diodes.

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FU WITH NONLINEAR FEEDBACK ON THE VOLTAGE/STRESS.

The parallel connection/communication, which is realized by the

inclusion/connection of resistor/resistance $R_{\text{нел}}$ between the output and input terminals of amplifier, is most widely used voltage feedback. Simplified circuit OF FU on the tubes with parallel nonlinear voltage feedback is shown on Fig. 43.

The resistor/resistance of nonlinear element/cell $R_{\text{нел}}$, whose value is changed with a change in the value of the voltage of signal, is the resistor/resistance to feedback. Consequently, with a change in the value of signal amount of feedback is changed, which produces change in input admittance of cascade/stage and transmission factor of the previous cascade/stage.

In connection with this for the realization of assigned FAX it is necessary to examine two series-connected cascades/stages, which let us conditionally name nonlinear pair. Transmission factor on the voltage/stress of the pair

$$\dot{K}_n = \dot{K}_1 \dot{K}_2, \quad (2-210)$$

where \dot{K}_1 and \dot{K}_2 - respectively the transmission factors of the first and second cascades/stages of pair.

For the vacuum-tube amplifier according to work [56]

$$\dot{K}_2 = \frac{S_2 - \dot{Y}_{a.02} - \dot{Y}_{\text{нел}}}{\dot{Y}_2 + \dot{Y}_{\text{нагр2}} + \dot{Y}_{a.02} + \dot{Y}_{\text{нел}}} \approx \frac{S_2 - \dot{Y}_{\text{нел}}}{\dot{Y}_{\text{нагр2}} + \dot{Y}_{\text{нел}}}, \quad (2-211)$$

since usually (especially for the pentodes) are fulfilled by

inequalities $\dot{Y}_{\text{max}} \gg \dot{Y}_{\text{a.o.}}$ and $\dot{Y}_{\text{marp}} \gg \dot{Y}_{\text{a.}}$

$$\dot{K}_1 = \frac{S - \dot{Y}_{\text{a.o.}}}{\dot{Y}_{\text{a.}} + \dot{Y}_{\text{marp1}} + \dot{Y}_{\text{a.o.}} + \dot{Y}_{\text{az}}} \approx \frac{S_1}{\dot{Y}_{\text{marp1}} + \dot{Y}_{\text{az}}}, \quad (2-212)$$

since for pentodes $S \gg \dot{Y}_{\text{a.o.}}$

Input admittance of the second cascade/stage

$$\dot{Y}_{\text{az2}} = \dot{Y}_{\text{a.}} + (\dot{Y}_{\text{a.o.}} + g_{\text{max}})(1 - \dot{K}_2).$$

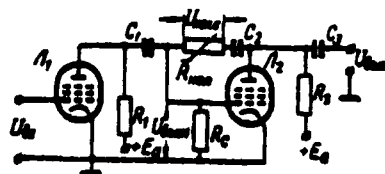


Fig. 43. Simplified circuit of FU with parallel nonlinear voltage feedback.

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Since $g_{\text{нел}} \gg Y_{\text{a.o}}$ and the second cascade/stage is loaded to the effective resistance (the aperiodic amplifier in the region of medium frequencies or selective amplifier at the resonance frequency), according to work [56]

$$g_{\text{вх2}} = g_0 + g_{\text{нел}}(1 + K_2).$$

With the work of tube without grid currents $g_0 \approx 0$. Then

$$g_{\text{вх2}} \approx g_{\text{нел}}(1 + K_2).$$

Substituting the value of K_1 and K_2 in (2-210), we obtain

$$K_{\text{н}} = \frac{S_1(S_2 - g_{\text{нел}})}{[g_1 + g_{\text{нел}}(1 + K_2)](g_2 + g_{\text{нел}})}, \quad (2-213)$$

where $g_1 = \frac{1}{R_1} = g_{\text{вхр1}}$; $g_2 = \frac{1}{R_2} = g_{\text{вхр2}}$ - are constant values.

With the work of pair in the linear conditions

$$K_{\text{н.л}} = \frac{S_{1\text{л}}(S_{2\text{л}} - g_{\text{нел.л}})}{[g_{1\text{л}} + g_{\text{нел.л}}(1 + K_{2\text{л}})](g_{2\text{л}} + g_{\text{нел.л}})}. \quad (2-214)$$

Taking into account expressions (2-18), (2-213), (2-214) and solving them relative to the conductivity of nonlinear element/cell

$g_{\text{нел}}$, we obtain

$$g_{\text{нел}} = \frac{S_{\text{н1}} \Phi(S_1) S_{\text{н2}} \Phi(S_2) - g_1 g_2 \varphi(z)}{K_{\text{н.н}} \varphi(z) (g_1 + g_2 + S_{\text{н2}} \Phi(S_2) + S_{\text{н1}} \Phi(S_1))}. \quad (2-215)$$

The voltage of signal on the nonlinear element/cell according to Fig. 43

$$U_{\text{нел.с}} = U_{\text{нмх}} - U_{\text{вх2}} = U_{\text{нмх}} - U_{\text{нмх1}}, \quad (2-216)$$

where $U_{\text{вх2}}$ - voltage on the input of the second cascade/stage, equal to the output voltage/stress of the first cascade/stage.

Normalizing voltage/stress $U_{\text{нел.с}}$, we obtain

$$z_{\text{нел}} = z - z_1, \quad (2-217)$$

where

$$z_1 = \frac{U_{\text{нмх1}}}{U_{\text{нмх.н}}} = \frac{K_1}{K_{\text{н.н}}} x. \quad (2-218)$$

Current through the nonlinear element/cell

$$I_{\text{нел.с}} = U_{\text{нел.с}} g_{\text{нел.с}} = U_{\text{нмх.н}} \left(z - \frac{K_1}{K_{\text{н.н}}} x \right) g_{\text{нел.с}}. \quad (2-219)$$

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For the transistor amplifier, assembled on the common-emitter connection, according to work [22] we have

$$K_1 = \frac{\dot{S}_1}{\dot{Y}_{\text{т1}} + \dot{Y}_{\text{нарп1}} + \dot{Y}_{\text{вх2}}} \approx \frac{\dot{S}_1}{\dot{Y}_{\text{нарп1}} + \dot{Y}_{\text{вх2}}}; \quad (2-220)$$

$$K_2 = \frac{\dot{S}_2}{\dot{Y}_{\text{т2}} + \dot{Y}_{\text{нарп2}}} \approx \frac{\dot{S}_2}{\dot{Y}_{\text{нарп2}}}, \quad (2-221)$$

since it is possible to consider that $\dot{Y}_{\text{вх2}} > \dot{Y}_{\text{т1}}$ and $\dot{Y}_{\text{нарп2}} > \dot{Y}_{\text{т2}}$. Input admittance of the second cascade/stage

$$\dot{Y}_{\text{вх2}} = \dot{Y}_2 - \dot{Y}_{\text{нарп}} K_2 = \dot{Y}_2 + \dot{S}_2 \frac{g_{\text{нел}}}{\dot{Y}_{\text{нарп2}}}.$$

Transmission factor of the pair

$$K_n = K_1 K_2 \frac{S_1 S_2}{\left(Y_{\text{нагр}1} + Y_1 + S_2 \frac{g_{\text{нел}}}{Y_{\text{нагр}2}} \right) Y_{\text{нагр}2}}. \quad (2-222)$$

Taking into account expression (2-118) and solving relatively $g_{\text{нел}}$, we obtain

$$g_{\text{нел.с}} = \frac{S_{1n} \Phi(S_1) S_{2n} \Phi(S_2) - Y_{\text{нагр}2} (Y_{\text{нагр}1} + Y_1) K_{\text{н.н}}(z)}{S_{2n} \Phi(S_2) K_{\text{н.н}}(z)}. \quad (2-223)$$

For aperiodic amplifier $Y_1 = g_1$, $Y_{\text{нагр}2} = g_{\text{нагр}2} = g_2 + \frac{1}{R_{\text{н}2}}$, $Y_{\text{нагр}1} = g_1 = \frac{1}{R_{\text{н}1}}$, for the resonance - $Y_{\text{нагр}1} \approx g_{\text{ос}1}$, $Y_{\text{нагр}2} = g_{\text{ос}2} + g_2$, where g_1 - the active component of input admittance of the first cascade/stage of the following pair.

Expressions (1-216)-(2-219) are valid also for the transistor amplifier.

From expressions (2-215) and (2-223) it is evident that for the realization of FAX with $b = \text{var}(\downarrow)$ is required the nonlinear element/cell with 1st type characteristics, and with $b = \text{var}(\uparrow)$ - with 2nd type characteristics.

During the calculation of dependence $g_{\text{нел.с}} = f(U_{\text{нел.с}})$ it is first necessary to calculate $z_{\text{нел}} = \varphi(z)$ and $g_{\text{нел.с}} = \varphi(z)$, and then for concrete/specific/actual values $K_{\text{н.н}}$, K_1 and $U_{\text{нел.с}}$ - curved $g_{\text{нел.с}} = f(U_{\text{нел.с}})$ or dynamic volt-ampere characteristic of nonlinear

element/cell.

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The character of curves $q_{\text{non. o}} = f(U_{\text{non. o}})$ depends on value $K_{\text{a. n}}$ and ratio of the factors of amplification of the first and second cascades/stages in linear conditions $\gamma = \frac{K_{2\text{a}}}{K_{1\text{a}}}$. The value of coefficient γ it is necessary to choose more than one. With value $\gamma < 1$, in the first place, the second cascade/stage is considerably overloaded by large signals, which is especially noticeable in the transistor amplifiers, and, in the second place, diagram can prove to be virtually impracticable from the point of view of obtaining the required amplitude characteristic of pair.

During the initial calculation of dependence $q_{\text{non. o}} = f(U_{\text{non. o}})$ possible to take into account a change in the slope/transconductance only of first amplifier instrument S_1 , and a change in slope/transconductance S_2 second UP can be taken into account, using the method of iterations.

It is necessary to note that the input resistance of the 1st cascade/stage of pair in the transistor amplifiers also noticeably is changed. For facilitating of tuning/adjusting and adjustment of logarithmic transistor amplifiers on the pairs it is expedient

between the pairs to include the untying emitter followers.

Besides the circuit solutions examined, functional amplifier can be made with the complicatedly nonlinear feedback. The results of experimental research and fundamental amplifier circuits with the nonlinear feedback on the transistors are given in work [8] and (Fig. 64, 73).

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Chapter 3.

FUNCTIONAL AMPLIFIERS WITH AUTOMATIC GAIN CONTROL.

§1. Classification and comparative evaluation of diagrams of ARU.

Classification of diagrams of ARU.

For obtaining FAKh in the selective and aperiodic amplifiers it is possible to apply ARU, accomplished with the help of the controlling voltage/stress. Examining an amplifier of any type in the form of active electrical network, it is not difficult to see that the transmission factor can be changed, changing:

- 1) the slope/transconductance of passage characteristic of UP;
- 2) equivalent resistance of load;
- 3) transmission factor in input or output circuit of

cascade/stage.

The slope/transconductance of passage characteristic of UP can be changed, changing the operating mode on the direct or alternating current (by change in the depth of negative feedback). The second and third methods can be carried out, applying the controlled attenuators.

In accordance with this let us note the following methods of the automatic adjustment of the transmission factor of cascade/stage; a change in the mode of operation of UP in the direct or alternating current (change in the depth of OOS); a change in equivalent resistance of load or transmission factor in the input (output) circuit of cascade/stage.

The possible circuit solutions of ARU are enumerated in Table 2.

In terms of the method of supplying the controlling voltage/stress of diagram of ARU it is possible to divide into the looped systems (Fig. 44a), when controlling voltage/stress U_v is supplied from the subsequent cascades/stages in previous (ARU back) and the system without the feedback (Fig. 44b), when the controlling voltage/stress is supplied from the previous cascades/stages into following (ARU forward). Any system of ARU encompasses the regulator,

which usually consists of detector (D), amplifier of regulator (UR) and filter (F).

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Regulator (R) under the action of output (Fig. 44a) or input (Fig. 44b) voltages develops controlling voltage/stress U_p , which enters the adjustable cascade/stage (RK). So that the regulator would operate/wear on the completely specific signal level, to it is given cutoff voltage U_s (voltage/stress of delay). Voltage/stress U_s can be given to the detector of ARU; to the amplifier of regulator and to the circuit of the controlling voltage/stress (to the further element/cell in the circuit of filter). In accordance with this

$$\left. \begin{aligned} U_p &= K_Y k_R (U_{sx, p} - U_s); \\ U_p &= K_Y (k_R U_{sx, p} - U_s); \\ U_p &= K_Y k_R U_{sx, p} - U_s. \end{aligned} \right\} (3-1)$$

where $U_{sx, p}$ - voltage of signal, which enters the input of regulator. For the system of ARU back $U_{sx, p} = U_{smx}$ for ARU forward - $U_{sx, p} = U_{sx}$; K_Y -transmission factor UR; k_R - transmission factor of detector.

System of ARU must operate/wear with the voltage of signal, which corresponds to beginning of FAKh. In this case

$U_{sx, p} = U_{sx, p, n} = (U_{sx} \text{ or } U_{smx, n})$. Then according to expressions (3-1)

$$\left. \begin{aligned} U_s &= U_{sx, p, n}; \\ U_s &= k_R U_{sx, p, n}; \\ U_s &= K_Y k_R U_{sx, p, n}. \end{aligned} \right\} (3-2)$$

System of ARU both with the reverse/inverse and without the feedback can have the diverse variants. The basic versions of system ARU with the feedback are given in Fig. 45:

1) local ARU, when circuit ARU contains only the one adjustable cascade/stage (Fig. 45a);

2) multistage ARU, when circuit of ARU contains two (Fig. 45b), three or n of cascades/stages (Fig. 45c, d);

3) crossed ARU through one (Fig. 45e), two or more than cascades/stages.

In turn, multistage ARU can be with n regulators (Fig. 45c) and with one regulator (Fig. 45d).

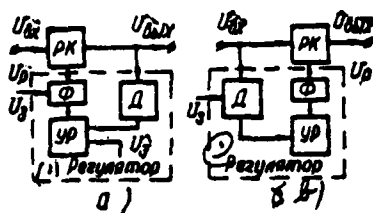


Fig. 44. Cascade/stage with the automatic gain control: a) ARU with the feedback (ARU back); b) ARU without the feedback (ARU forward).

Key: (1). Regulator.

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On the analogous block diagrams can be made the systems of ARU forward.

In Fig. 45 filters in the circuits ARU are omitted, since during the analysis of amplitude properties of FU we will examine the static behavior of the work of amplifier.

In terms of operating speed systems of ARU divided into the inertial ones (IARU) and moving rapidly (BARU). Speed of response of system ARU is determined in essence by the elements/cells of filter. The dynamic work of systems ARU is sufficiently fully examined in works [36, 59, 65.70, 77, 93].

The form of A_{Kh} of amplifier with ARU in essence is determined by the transmission factor of regulator. By changing the transmission factor of amplifier in the circuit ARU, it is possible to obtain various forms of A_{Kh} of cascade/stage and amplifier as a whole. It is obvious that the amplifier with $F A_{Kh}$ with $b = \text{var}(\uparrow)$ can stably work only with ARU forward, i.e., without feedback. System ARU with the feedback can be used for obtaining $F A_{Kh}$ with $b = \text{var}(\downarrow)$. However, during the use of a system of ARU with the feedback the depth of the adjustment of cascade/stage (amplifier) cannot be obtained more than some specific value and, consequently, also the inclination/slope of A_{Kh} of cascade/stage (amplifier) with an increase in the transmission factor K_p of amplifier in the circuit ARU cannot be obtained more than completely determined.

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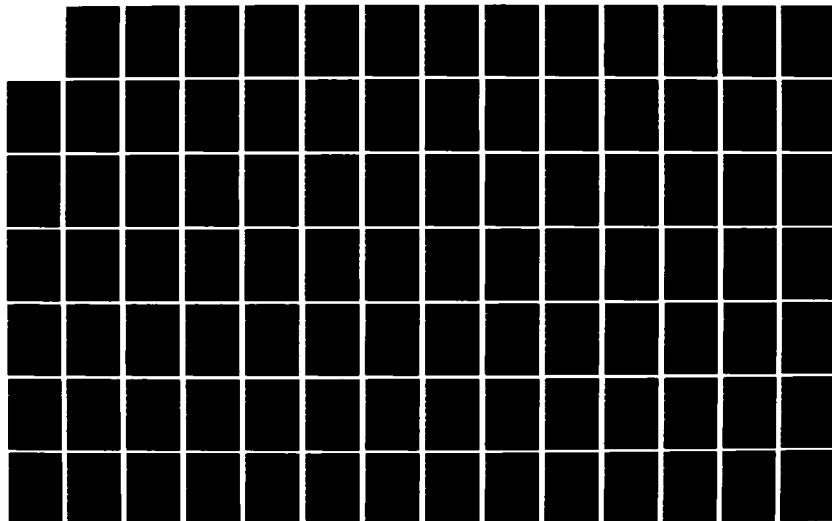
FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC
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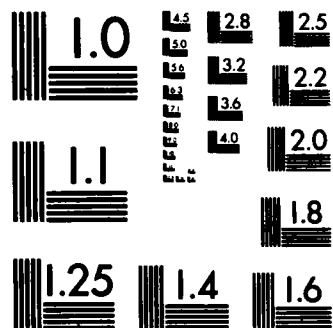
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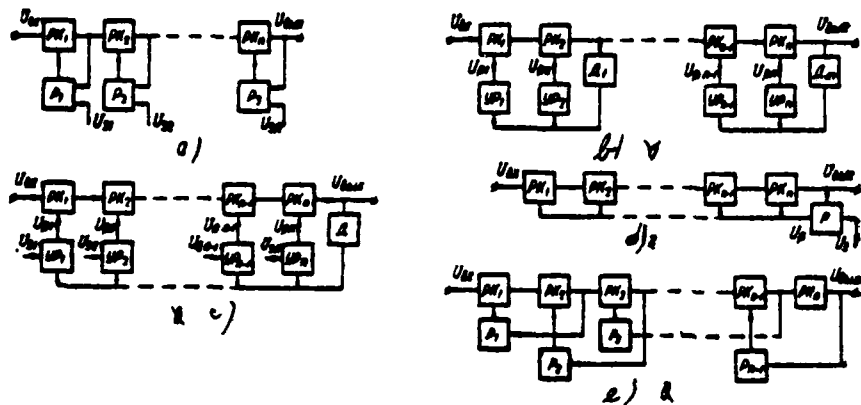


Fig. 45. The basic versions of system ARU with the feedback.

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Region 1, in which can be located AKh the cascade/stage (amplifier), included by ARU with feedback, in Fig. 46 it is shaded. With an increase in the transmission factor K_p (K_p^I , K_p^{II} , K_p^{III} and so forth) AKh cascade/stage approaches an axis of abscissas. But in the control systems with the feedback it cannot go in parallel to the axis of abscissas. Consequently, with the aid of ARU with feedback it is not possible to realize ideal limitation with the differential transmission factor $b=0$. The system of ARU without feedback is deprived of this deficiency/lack, applying which it is possible with the appropriate transmission factor K_p to obtain the characteristic of a cascade/stage of any type (curves 1, 3, 4 in Fig. 46). If, beginning from some level of input signal U_{ex} , the amplification of

cascade/stage is changed according to the law

$$K = \frac{m}{U_{ax}},$$

then with $m=1$ the characteristic of cascade/stage will go in parallel to the axis of abscissas (curve 3 in Fig. 46), for $m<1$ the characteristic will have slope to the axis of abscissas (curve 4) and with $m>1$ voltage at the output of the adjustable cascade/stage will grow with an increase in the signal (curve 2). The necessary value of coefficient of m can be obtained during the appropriate selection of transmission factor K_p . This is explained by the fact that in the system in question the amplifier in the circuit of ARU is not included by feedback.

In the system of ARU without feedback the time constant of functioning can be selected very small. Consequently, this system of gain control can be used for obtaining FAKh in the pulse aperiodic and selective amplifiers, intended for amplifying the impulses/momenta/pulses of short duration.

Deficiencies/lacks in the system of ARU without feedback are: the need for guarantee in the circuit of ARU of large amplification, criticality to a change in the ambient temperature and to the exchange of network elements of ARU, especially in the transistor amplifiers. Therefore systems of ARU with the feedback are more widespread.

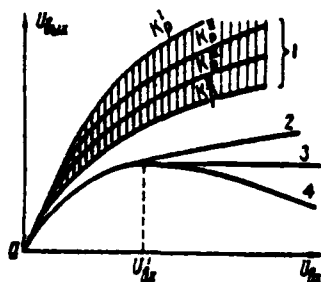


Fig. 46. Amplitude characteristics of amplifier with ARU.

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SHORT COMPARATIVE EVALUATION OF SYSTEMS OF ARU.

In the vacuum-tube amplifiers the triodes, tetrodes and pentodes can be used. If a tube of the pentode type is used, then its slope/transconductance can be changed, changing voltage/stress on one of the networks: control, shielding or pentode. With decrease of voltage on control electrode operating point it is displaced from the straight portion of the passage characteristic of tube to the curvilinear. Because of this with an increase in the harmonic signal the nonlinear distortions of the reinforced signal grow. Therefore ARU on control electrode of tube one should apply in the aperiodic amplifiers of video pulses and in the selective amplifiers of the harmonic oscillations, in load of which the fundamental harmonic of the reinforced signal is selected.

A change in the input capacitance with a change in the bias voltage is a deficiency/lack of ARU on control electrode of tube.

However, a change in the input capacitance of tube can be considerably decreased, join up of the cathode of the resistor/resistance of feedback $R_{o.o.}$. In this case input capacitance according to expression (2-209)

$$C_{ex.o.o.} = \frac{C_{ex.o.} \Phi_c(U_p)}{1 + S_{\mu} \Phi_s(U_p) R_{o.o.}},$$

where $\Phi_c(U_p)$ and $\Phi_s(U_p)$ - decreasing functions with the increase of the negative controlling voltage/stress; $C_{ex.o.}$ - initial value of input capacitance.

It is the condition of the stabilization of capacity/capacitance C_{ex} with change U_p

$$\frac{dC_{ex}}{dU_p} = 0.$$

Solving this condition relative to resistor/resistance $R_{o.o.}$ we obtain

$$R_{o.o.} = \frac{\Phi_c'}{S_{\mu}(\Phi_s' \Phi_c - \Phi_c' \Phi_s)}.$$

For calculation $R_{o.o.}$ it is necessary to know the analytical expressions of functions Φ_s and Φ_c . Expressions for Φ_s are given in Table 9.

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Expression for ϕ_0 can be determined according to the experimental data. Resistor/resistance $R_{0.}$ must be calculated for the average/mean value of the controlling voltage/stress.

Applying the method of stabilization examined, it is necessary to remember that upon the inclusion/connection of resistor/resistance $R_{0.}$ the initial transmission factor of cascade/stage is reduced.

The advantage of diagram of ARU on control electrode is that for the adjustment a comparatively small control voltage is required and power virtually does not consume the circuit of adjustment. In the circuit of ARU the gain amplifier can be used.

With the decrease of voltage/stress on the screen grid the passage characteristic of tube is shifted into the region of positive voltages/stresses on control electrode. With the decrease of slope/transconductance the extent of the linear section of passage characteristic simultaneously is reduced. During the gain control on the screen grid with the increase of signal nonlinear distortions grow and upon reaching/achievement of the signal amplitude of a comparatively small level the circuital currents of control electrode of tube appear, which leads to the shunting of the load of the

previous cascade/stage.

For changing the slope/transconductance of tube within considerable limits the large controlling voltage/stress, congruent in the value with the voltage/stress on the screen grid, is necessary.

Due to the enumerated deficiencies/lacks the diagram of ARU on the screen grid in FU in practice is not applied.

Special attention deserves the adjustment of the slope/transconductance of tube by a change in the voltage/stress on the pentode grid. The characteristics of a tube of the type 6Zh9P (on the anode, the cathode and the screen grid) depending on voltage/stress on pentode grid U_n for four values of bias voltage U_0 on control electrode are given in Fig. 47a, and the plate-grid characteristics at different values U_n - in Fig. 47b. From the characteristics given in the figure it is evident that the dependence of the anode current of tube on U_n is almost linear in the considerable range of change U_n ; anodically - the grid characteristics of tube with an increase in negative voltage/stress U_n are forced against the axis of abscissas; the extent of the linear section of characteristic in this case not only is retained, but even increases.

In view of the properties of tube examined the gain control in the cascade/stage is realized without the nonlinear distortions with a change of the amplitude of harmonic signal in ShDD.

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Thus, gain control by a change in the voltage/stress on the pentode grid it is necessary to apply in FU of harmonic signals with small nonlinear distortions, in particular in the aperiodic low-frequency amplifiers, used in the measuring equipment.

The advantage of the method of gain control in question is a small change of the input capacitance of tube in the alignment procedure, by deficiency/lack - need for supply to the pentode grid of the considerable negative controlling voltage/stress.

If as UP transistor is used, then the amplification of cascade/stage by a change in the operating mode in the direct current can be regulated:

1) by a change in the current of the emitter of transistor under the effect of the controlling voltages/stresses U_p on the current emitter (positive governing) and base current (indirect adjustment);

2) by a change in the collector voltage of the transistor of the adjustable cascade/stage under the effect of voltage/stress of ARU on the collector voltage/stress (positive governing), the base current (indirect adjustment) and the current of emitter (indirect adjustment);

3) by a simultaneous change in the current of emitter and collector voltage/stress under effect U_p .

Systems of ARU, based on a change in the mode/conditions of transistors in the direct-current, have a series/row of specific special features/peculiarities in comparison with the electron-tube systems of ARU. Their fundamental difference lies in the fact that complete input and output conductances of transistors are very great and are changed within considerable limits in the control [27, 42, 66, 69, 109].

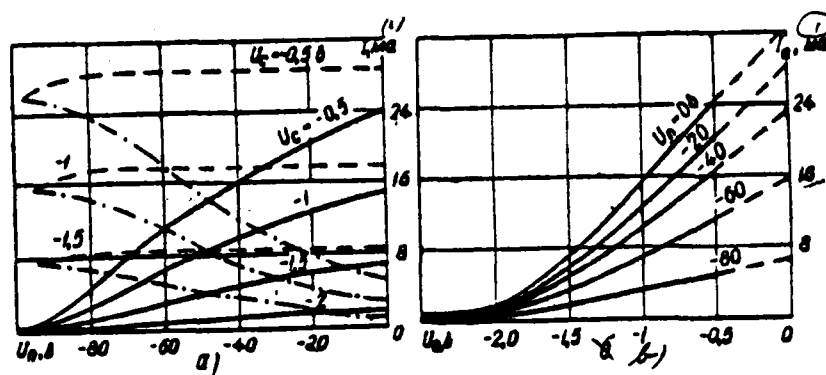


Fig. 47. Characteristics of a tube of the type 6Zh9P.

Key: (1). mA.

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The corollary of this is the decrease of the depth of adjustment; a change in the frequency of tuning/adjusting, passband, energy factor of the oscillatory circuit and phase of output signal in the control; the consumption of considerable power from the regulator, especially during the positive governing in the current of emitter and in the voltage/stress of collector/receptacle.

Let us consider the diagrams of adjustment from the point of view of the efficiency of adjustment. Analyzing the input and passage characteristics of transistor [8], we see that the parameters of transistor and, consequently, also the transmission factor of

cascade/stage considerably are changed for the voltages/stresses on the collector/receptacle less than 1 V. When $U_m > 1$ V, especially in the diagram with OB, change U_m within large limits barely affects the parameters of transistor and its amplifier properties. This phenomenon is analogous with a change in the anode voltage of vacuum lamp of the type pentode.

For low values U_m when transistor works in the saturation mode, characteristic of transistor they converge and they change inclination/slope, the amplifier properties of transistor are reduced.

In the diagram with OE the range of change U_m in which a change of the parameters of transistor is perceptible, is considerably more than in the diagram with OB. Therefore it is possible to obtain the required amplitude characteristic of the adjustable cascade/stage.

With the low currents of base and, consequently, also the currents of emitter, begins the cutoff conditions and the amplification of transistor sharply falls.

Voltage/stress on the collector/receptacle with the insignificant expenditure of power can be regulated by indirect method, i.e., by the supply of the controlling voltage/stress into

the circuit of base. In this case in the circuit of collector/receptacle it is necessary to include/connect the sufficiently high resistor/resistance (units of kilohms) on direct current [8]. This method of control can be applied only with the low-voltage feed, which is not always justified.

Rather simple and efficient is the method of gain control a change in emitter current [8]. In this case the controlling voltage/stress can be supplied both into the emitter circuit and into the circuit of base. With the direct method of control of the adjuster considerable power is required.

For the realization of FAKh in ShDD it is necessary to change I_{*} and, accordingly, I_{*} over wide limits. However, for a number of reasons it is necessary to limit these limits.

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The maximum value of collector cutoff current I_{*m} in the absence of control signal is chosen:

1) from the condition of the maximum permissible power, scattered on the collector/receptacle of load. Even the insignificant heating of transistors due to action I_{*} can cause instability of ARU.

Based on this condition, value I_k for the transistors of types P401, P402, P403, P411 and others must be chosen in the limits to 10 mA;

2) from the condition of obtaining the minimum power of control;

3) on the basis of reference data of the parameters of the hf/VCh transistors, which are given for standard mode/conditions $I_{k, \text{max}} = 5 \text{ mA}$;

4) from the minimally permissible current of collector/receptacle.

With decrease of I_k with the decrease of conductivity g_{11} conductivities g_{11} and g_{12} are reduced also. With what that value I_k is observed maximum relation $\frac{g_{11}}{g_{11} + g_{12}}$ in which it is possible to obtain the maximum transmission factor of cascade/stage on the voltage/stress.

With the low currents of collector/receptacle and the large signals the series/row of the undesirable phenomena appears, basic from which are [34]:

1) a deterioration in the temperature stability of the adjustable cascades/stages as a result of the increase of the effect

of the non-controlled current of collector/receptacle I'_{K0} on the work of diagram. Value I'_{K0} is considerably more than I_{K0} for the common-base circuit, given in the manuals;

2) the possibility of the transition of transistor into the mode/conditions of self-detection, during which constant component I_K increases with the increase of signal and the work of system of ARU is broken ¹.

FOOTNOTE ¹. The mode/conditions of self-detection is used as useful phenomenon for the realization of gain control chapter 3, § 2.

ENDFOOTNOTE.

For the realization of assigned FAKh it is necessary to consider the effect of self-detection or to remove it;

3) a sharp increase in the nonlinear distortions, what is undesirable phenomenon during the amplification of the modulated vibrations. However, in FU, the intended for the amplification pulse signals, the phenomenon indicated is not substantial.

For the high-frequency transistors of types P401-P402, P411, P416 and others the small effect of gain control by base current (emitter) is achieved by a change in the base current from 50-100 μA

to zero, which corresponds to a change in the current of emitter from 3-10 mA to zero.

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If in the circuit of collector/receptacle load on the direct current (load resistance/resistor in the low-frequency amplifiers, the resistor/resistance of decoupling filter in the tuned amplifiers) is connected, then with reduction in current of emitter and, consequently, also the current of collector/receptacle, in the control voltage/stress on the collector/receptacle increases, in consequence of which the amplifier properties of transistor are improved and the effect of control descends. Therefore in the resonance cascades/stages, adjusted by the current of emitter, it is necessary to apply series feed on the circuit of collector/receptacle and decoupling filters of the type LC.

Let us consider the effect of a change in the parameters of transistors for the resonance frequency and the passband of amplifier. Let us assume that the adjustable transistor is included in multistage amplifier and has respectively at the input input oscillatory circuit and at the output - output circuit. If transistor is connected to these ducts/contours incompletely with the transformation ratios m_1 and m_2 , then input and output resistance

introduced into the ducts/contours and capacities/capacitances of transistors are respectively reduced into m^1 , and m^2 , times.

According to works [6, 27, 42, 69] with reduction in current of emitter input resistance $R_{ex. in}$ introduced into the duct/contour, increases, and the value of introduced input capacitance $C_{ex. in}$ is reduced, as a result of which input circuit is disturbed/detuned in the direction of higher frequencies, and passband becomes narrow. The contraction of the passband of duct/contour is caused by the fact that a relative increase in the general/common/total resonance resistor/resistance of R_0 of duct/contour due to increase $R_{ex. in}$ is considerably more than the relative decrease of total capacitance C_0 of the duct/contour, caused by decrease $C_{ex. in}$.

Therefore product $C_0 R_0$ with decrease I_0 increases, and passband - is reduced.

Analogous phenomenon is observed also with a change in the output parameters of transistor, caused by change I_0 . As a result of an increase in that introduced into the duct/contour of resistor/resistance R_{out} the passband becomes narrow, and with comparatively small decrease C_{out} output circuit is disturbed/detuned in the direction of higher frequencies. Although percentage change in values R_{out} and C_{out} is less than R_{ex} and C_{ex} .

taking into account that the output circuit frequently is connected completely to the collector/receptacle of transistor, the effect of change R_{bx} and C_{bx} cannot be disregarded.

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During the gain control by the decrease of voltage/stress on the collector/receptacle resistors/resistances R_{bx} and R_{max} are reduced, and C_{bx} and C_{max} increase, as a result of which passband of input and output circuits they are widened and ducts/contours are disturbed/detuned in the direction of lower frequencies. Thus, during the adjustment with voltage/stress on the collector/receptacle the drift of resonance frequency and a change in the passband of ducts/contours by nature are opposite to the character of a change in these parameters at the adjustment by the current of emitter.

If we use combined gain control on the current of emitter and the voltage/stress of collector/receptacle, then it is possible to obtain the higher stability of the passband and resonance frequency, than in the diagrams with the adjustment only by the current of emitter or with voltage/stress on the collector/receptacle. During the combined gain control for input circuit increase/growth $R_{\text{bx. in}}$ and decrease $C_{\text{bx. in}}$ with decrease I_e are partially compensated by decrease $R_{\text{bx. in}}$ and increase $C_{\text{bx. in}}$ with decrease U_{bx} . Analogous phenomenon is

observed also for the output circuit. Since parameters R_{out} , R_{out} , C_{out} and C_{out} nonlinearly depend on I_b and U_{be} . During the appropriate selection of the initial point of control on U_{be} or I_b it is possible to ensure this compensation to a greater or lesser extent.

From the point of view of the frequency stability of the best the diagram of simultaneous gain control on the current of emitter and the voltage/stress of collector/receptacle is.

The schematic of the tuned amplifier, adjusted by simultaneous ones changing the current of emitter and voltage/stress on the collector/receptacle, and its amplitude characteristics are given in work [8].

During the gain control by a change in the mode/conditions of transistor in the direct current the reverse/inverse conductivity Y_{12} of transistor is changed. Upon the start of transistor on the diagram with OE and during the adjustment by base current (by voltage/stress U_{be}) conductivity Y_{12} insignificantly depends on the current strength. Therefore in the selective amplifiers of radio frequency it is possible to use the diagrams of neutralization.

Thus, in the implementation of FAKh in the transistorized amplifiers it is most expedient transistors to switch on on the

diagram with OE and to apply gain control by change U_{be} (base current).

During the creation of FU with the adjustment by a change in the mode/conditions of transistor in the direct current appear the further difficulties, caused by the essential dependence of the parameters of transistor on changes in the ambient temperature.

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Inverse current of the collector/receptacle of the transition is the main factor, which are determining the temperature dependence of the work of transistor,

$$I_{\text{KO}}(t) = I_{\text{KO}} e^{\gamma t},$$

where $\gamma = (0.06 \div 0.08) \pm 0.01 \text{ deg}^{-1}$ - temperature coefficient determined by the properties of semiconductor material.

For the stabilization of the current of collector/receptacle are applied different diagrams of temperature stabilization, in particular, with the help of temperature-independent resistors/resistances and the special thermistors, connected with the circuit of the application of voltage of bias/displacement on the base of transistor. As the temperature elements/cells the thermistors of types KMT-4, MMT-4 and other, germanium or silicon diodes in the

dependence on the type of transistor (p-n-p or n-p-n) can be used. It is more expedient to apply semiconductor diodes, since they have thermal inertia identical to the transistors, while the thermistors this property do not possess, in consequence of which it is necessary to experimentally select network elements of the application of voltage of bias/displacement.

During the temperature stabilization of mode/conditions with the help of the linear resistor/resistance of feedback R_* connected with the emitter circuit, the range of adjustment is reduced with increase R_* [48] with the given controlling voltage/stress. Consequently, for obtaining the required effect of gain control with increase R_* it is necessary to increase the voltage/stress of adjustment.

For eliminating the deficiency/lack indicated it is proposed to preserve the chain/network of self-bias in the emitter circuit, but emitter to simultaneously connect with the output of filter ARU through the semiconductor diode, ensuring the increased depth adjustment of ARU and retaining the required mode/conditions of the work of transistor [45]. In the patent indicated are given the diverse variants of the fulfillment of diagram of ARU, moreover in all versions transistor is connected on the diagram with OB.

The special features/peculiarities of the construction of

channel of ARU are examined in works [27, 100, 101, 108].

For increasing qualitative indices of FU on the transistors are applied the methods and the diagrams of the adjustments, which earlier were not applied in the vacuum-tube amplifiers: the controlled attenuators (dividers) and shunts, the adjustable feedback and so forth [80, 81, 85, 97].

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§2. Functional amplifiers with ARU with change in the mode of operation of UP on direct current.

ARU on control electrode of UP.

For the realization of assigned FAKh it is necessary to the amplifier instruments to supply U_p , that is changed in the dynamic range according to the law, which is determined by type of FAKh, form of passage characteristic of UP and by type of system of ARU.

For developing the system of ARU of functional amplifier it is necessary to know dependence $U_p = f(U_{ax.p})$, from which it is possible to calculate the required character of a change in the transmission factor $K_p = f(U_{ax.p})$ and to fit the appropriate component elements of

where U_{cm} - bias voltage.

$$U_{cm} = -U_{a.0} - U_p \quad (3-7)$$

For the tube

For the transistor of type p-n-p

$$U_{cm} = -U_{a.0} + U_p \quad (3-8)$$

where $U_{a.0}$ - initial bias voltage.

According to expression (2-20) for the synthesis of FU with ARU it is necessary to have a dependence $\Phi(S) = f(U_{cm}, U_m)$, which is the function of the bias voltage and amplitude of input signal. Such dependences for the most widely used approximations of passage characteristics of UP are given in Table 9.

Let us examine in more detail the case of piecewise-parabolic approximation, which with degree of $p=1$ is converted into the piecewise-linear (polygonal). In this case (Fig. 48)

$$\cos \theta = \frac{U_{cm} - U_0}{U_{a.m}}, \quad (3-9)$$

where U_0 - cutoff voltage of anode current for the tube.

FOOTNOTE ¹. Further index m is omitted. ENDFOOTNOTE.

Bias voltage

$$U_{cm} = U_{e. n} + U_p = U_0 - U_{ex. n} + U_p \quad (3-10)$$

since $U_{e. n} = U_0 - U_{ex. n}$. Substituting $U_{e. n}$ into expression (3-9), we obtain

$$\cos \theta = \frac{-U_{ex. n} + U_p}{U_{ex. n}} \quad (3-11)$$

Whence

$$U_p = U_{ex. n} (x \cos \theta - 1). \quad (3-12)$$

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With the degree of parabola $p=2$ (Fig. 48a) the initial section of characteristic is approximated more accurately. Therefore for more precise calculation U_p with the low signals should be used approximation with $p=2$, and with the large signals - with $p=1$.

Let us consider the case of the realization of the most widely used linear and logarithmic amplifiers of harmonic oscillations.

Linear amplifier.

For the linear amplifier according to expression (2-20) must be implemented the equality

$$\Phi(S) = \Phi(Y). \quad (3-13)$$

Then with the square-law characteristic of UP controlling voltage/stress $U_p = 0$.

With the exponential approximation

$$\frac{e^{-U_p} I_1(\gamma x U_{sx, u})}{x I_1(\gamma U_{sx, u})} = \Phi(Y).$$

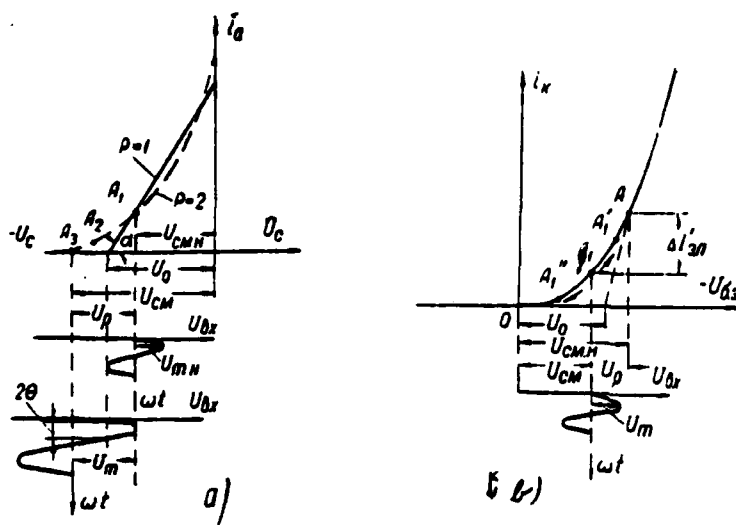


Fig. 48. Piecewise-parabolic approximation of the characteristics of the amplifier instruments: a) tube; b) transistor.

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Whence

$$|U_p| = \pm \frac{1}{\gamma} \ln \frac{2J_1(\gamma U_{ex, u}) \Phi(\gamma)}{J_1(\gamma U_{ex, u})}, \quad (3-14)$$

where $I_1(\gamma U_{ex}) = -jJ_1(j\gamma U_{ex})$ - modulus/module of the complex variable function first order.

For the transistors of type p-n-p U_p it is taken positive polarity, for tubes and transistors of the type n-p-n - negative.

In the case of n-cascade amplifier, made on the block diagram

(Fig. 45a), for the i cascade/stage

$$U_{pi} = \pm \frac{1}{\gamma} \ln \frac{I_1(\gamma U_{ex, n}) \Phi(Y)}{I_1(\gamma U_{ex, n-1})}, \quad (3-15)$$

where $U_{ex, n-1} = U_{ex, n} K_n^{i-1}$; $U_{ex, n}$ - the input voltage of the amplifier, with which the system of ARU operates/wears; K_n - transmission factor of cascade/stage. In this case it is assumed that all cascades/stages are identical and have identical transmission factor.

The system of ARU of the i -th cascade/stage must operate/wear with output potential of the i -th cascade/stage

$$U_{ex, pi} = U_{max, i} = K^i U_{ex, n}$$

The transmission factor of regulator during the supplying of the voltage/stress of delay on the amplifier of regulator according to expressions (3-1) and (3-3) in the alignment procedure must be changed

$$K_{pi} = \frac{U_{pi}}{U_{ex, pi}} = \frac{1}{\gamma K^i U_{ex, n} (i-1)} \cdot \ln \frac{I_1(\gamma U_{ex, n})}{I_1(\gamma U_{ex, n-1})}. \quad (3-16)$$

Generalized curves $\beta=f(\alpha)$ and $\xi=f(\alpha)$ on the assumption that load cascade constant, i.e., $\Phi(Y)=1$, are given in Fig. 49: Curves can be used for calculation U_p and K_p for different amplifier elements/cells and values of the transmission factor of cascade/stage K .

Curves $U_p=f(U_{ex})$ and $K_p=f(U_{ex})$ for the different values γ and are given in Fig. 50. From the figures it is evident that dependences $U_p=f(U_{ex})$ in form call to mind the transfer characteristics of

detector, in consequence of which it is possible to realize linear amplification. It is obvious that to combine the objective parameter of regulator $U_p = f(U_{ex.p})$ with that required is possible: along the axis of ordinates by a change in the transmission factor $K_{p,\gamma}$ of the amplifier of regulator, along the axis of abscissas - by change in voltages/stresses U_0 and $U_{ex.p}$ (latter is realized with the help of the divider, connected at the input of detector).

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Let us consider the possible dynamic range of the linear AKh of the amplifier, made on the high-frequency transistors of types P416-P417, for which $\gamma = (10-15)$.

It is virtually established/installed, that the maximum input surge voltage of radio frequency on the input of the cascade/stage, made on the transistors of types P416-P417, must not exceed 0.5-1 V. If we take $U_{ex. max} = 1$ V, then when $\gamma = 10$ $a_{max} = 10$; when $\gamma = 15$ $a_{max} = 15$. If we take the transmission factor of cascade/stage $K_n = 10$, then with $\gamma = 10$ we will obtain values a_{max} for the i^{th} cascade/stage of four-stage amplifier, given in Table 10.

From the table and the figures it is evident that in the first and the secondly cascades/stages a gain control in practice realized

must not be. However, the dynamic range of amplifier is very small. Therefore should be performed the amplifier, which consists of one or maximum of two adjustable cascades/stages.

Experimental data with an accuracy (in limits of 10-15%) sufficient for the practice coincide with the calculated ones.

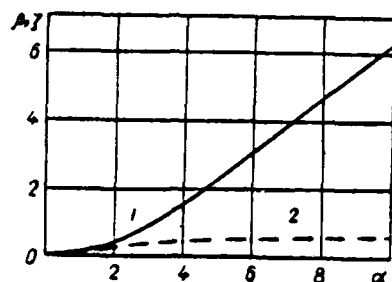


Fig. 49.

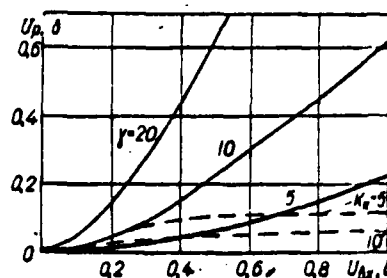


Fig. 50.

Fig. 49. Generalized dependences: 1 - $\beta=f(\alpha)$; 2 - $\zeta=f(\alpha)$.

Fig. 50. Curves of dependences for linear amplifier with ARU:

— $U_p = f(U_{BX})$; --- $\kappa_p = f(U_{BX})$.

Table 10.

	4	3	2	1
α_{MANO}	10	1	0,1	0,01
$U_{BX, MANO} \cdot \delta$	1	0,1	0,01	0,001

On the basis of theoretical and experimental studies it is possible to do the following conclusions:

1) with the help of ARU in the single-stage transistor amplifier linear AKh can be obtained to input voltage on the order of 0.5-1 V;

2) with an increase in the number of adjustable cascades/stages the upper limit of the dynamic range of the linear AKh of amplifier is reduced.

Logarithmic amplifier.

With square-law characteristic UP according to expression (2-20) and Table 9 for the amplifier, made on the block diagram (Fig. 45),

$$\frac{a_1 - 2a_2(U_{c.n} + U_p)}{a_1 - 2a_2U_{c.n}} = \frac{z}{s} \Phi(Y). \quad (3-17)$$

Solving equation relative to voltage/stress U_p , we obtain

$$U_p = \left(a_1 - A \frac{z}{s} \right) \frac{1}{2a_2} - U_{c.n}, \quad (3-18)$$

where

$$A = (a_1 - 2a_2U_{c.n}) \Phi(Y).$$

Expression (3-18) is correct for any FU with the substitution of the corresponding values z .

As a result of the calculations conducted and experimental research it is established/installed, that with the square-law characteristic of amplifier instrument with the help of ARU back due to a small adjustable section of characteristic UP successive work of cascades/stages and, consequently, also precise LAKh can be realized

in the two-stage maximum in the three-stage amplifier.

With the help of the system of ARU to in practice realize the forward successive work of cascades/stages is very difficult, since in this case the transmission factor of regulator K_p in the alignment procedure must always be changed according to the complicated law in the direction of decrease.

To considerably more easily realize LAKh in the multistage amplifier, made on the block diagram in Fig. 45, in which the amplification is regulated simultaneously in all cascades/stages.

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In this case the running transmission factor of cascade/stage must be described by expression (1-128), calibrated AKh - by expression (1-132), and the controlling voltage/stress

$$U_p = \left(a_1 - A \sqrt{\frac{Z}{X}} \right) \frac{1}{2a_1} - U_{c.c.} \quad (3-19)$$

where X and Z - standardized/normalized input and output voltages of amplifier.

Curves $U_p = f(Z)$, calculated by formula (3-19) for different number of adjustable cascades/stages n and on the assumption that $\phi(Y) = 1$,

are given in Fig. 51. During the calculation the approximation of the characteristic of tube 6Zh1P [12]

$$I_a = 20 + 9e_1 + e_2; U_{a. n} = -1 e; a = 1.$$

is accepted. From the curves in Fig. 51 it is evident that with an increase in the number of adjustable cascades/stages desired value U_p is reduced and dependence $U_p = f(Z)$ approaches linear, which indicates the possibility of the practical realization of precise LAKh in ShDD. For the diagram for Fig. 45d, transmission factor K_p , one should design from the formula

$$K_p = \frac{U_p}{U_{\text{max}}} = \frac{U_p}{U_{\text{max. n}} Z}. \quad (3-20)$$

As we see, for the realization of precise LAKh with the help of ARU back with the simultaneous adjustment in all cascades/stages with the square-law characteristic of UP it is necessary to take a number of adjustable cascades/stages not less than 4-5.

With polygonal approximation ($p=1$) with the strictly successive work of stages of the amplifier, made on the diagram in Fig. 45a

$$\frac{1}{\pi} \left(\theta - \frac{1}{2} \sin 2\theta \right) = \frac{\pi}{2} \phi(Y). \quad (3-21)$$

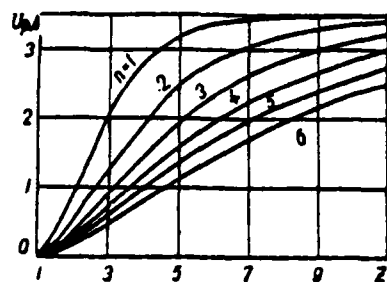


Fig. 51. Curves of the dependence of the controlling voltage/stress with different number of adjustable cascades/stages n .

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Expression (3-21) is transcendental and can be solved graphically. After equation (3-21) is solved relative to θ , from formula (3-12) the dependence

$$\eta = \frac{U_p}{U_{\text{max}}} = f(z) \text{ or } U_p = f(z).$$

is calculated. The normalized curves $\eta=f(z)$, designed for the different values of coefficient of a , $d=10$ and when $\Phi(Y)=1$, are given in Fig. 52. From the curves it is evident that the dependence $\eta=f(z)$ is first nonlinear (with the work of cascade/stage in the logarithmic mode/conditions) and then - linear (with the work of cascade/stage in the quasi-linear mode/conditions).

It is easy to obtain required dependence $U_p = f(U_{\text{ex. p}})$ with the help of the regulator, which consists of the linear dc amplifier and detector with the variable transmission factor with the low signals.

To considerably with more difficulty realize precise LAKh during the simultaneous gain control in all cascades/stages. In this case

$$\frac{1}{\pi} \left(\theta - \frac{1}{2} \sin \theta \right) = \sqrt[n]{\frac{U}{U_0}}. \quad (3-22)$$

The voltage/stress of adjustment for the i cascade/stage

$$U_{pi} = U_{\text{ex}i} \cos \theta + U_{\text{ex. st}i} \quad (3-23)$$

where $U_{\text{ex}i} = x U_{\text{ex. st}i} K_n^{1-i} \left(\frac{z}{x} \right)^{\frac{i-1}{n}}$ - voltage of signal on the input of the i -th cascade/stage.

From expression (3-23) it is evident that the value and the law of a change in the controlling voltage/stress are different for different cascades/stages, that it is possible to realize only in the amplifier, made on the block diagram (Fig. 45c).

With the approximation of characteristic UP by exponential curve and the successive work of nonlinear cascades/stages (Fig. 45a)

$$U_{pi} = \frac{1}{\gamma} \ln \frac{I_1(\gamma U_{\text{ex. st}i})}{I_1(\gamma U_{\text{ex. st}i})}. \quad (3-24)$$

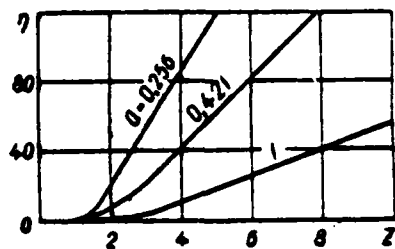


Fig. 52. Standardized/normalized curved with the polygonal approximation characteristics of UP.

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The generalized curves $\beta=f(z)$ for different values $\alpha_n = \gamma U_{\text{ex. n}}$ with $a=1$, $K_n=10$ and $\Phi(Y)=1$ are given in Fig. 53. From the curves it is evident that the adjustment for values $\alpha_n = 0.1-0.2$ is actually realized, when dependence $\beta=f(z)$ at values of $z=1-3.3$ (which corresponds to logarithmic mode/conditions) has nonlinear dependence, and when $z \gg 3.3$ - linear. When $\alpha_n > 0.2$, the function $\beta=f(z)$ approaches exponential, which indicates the possibility of realization of LAKh with the help of ARU forward with the proportional regulator, when $\alpha_n < 0.1$ - dependence $\beta=f(z)$ has complicated character.

Consequently, in the implementation of LAKh with the help of ARU back necessary to choose value $\alpha_n = 0.1-0.2$; in the implementation of LAKh with the help of ARU forward - to take value $\alpha_n > 0.3$.

If we take $\gamma=15$, $K_n=10$ and $\alpha_n=0.1$, then $U_{\text{ex. n1}} = 6$ mV. Then the cutoff voltage, which must be fed to the detector of regulator,

$$U_0 = U_{\text{det. n1}} = U_{\text{ex. n1}} K_n = 60 \text{ mV.}$$

A successive work of the adjustable cascades/stages also can be

carried out in the amplifier, made on the block diagram (Fig. 45c), if we to the detectors of regulators feed the cutoff voltages:

$$U_{pi} = \dot{U}_{max. n} [(i-1) \ln d + 1], \quad (3-25)$$

where i - the reference number of cascade/stage with the reading from the end/lead.

During the simultaneous gain control, when the identical law of a change in the transmission factors of cascades/stages is satisfied, variable voltage for the i cascade/stage of an n -stage amplifier (with the reading from the first cascade/stage)

$$U_{pi} = \frac{1}{\gamma} \ln \frac{I_1(\gamma U_{max. n})}{\sqrt[n]{X^{n-1} 2^i I_1(\gamma U_{max. n})}}. \quad (3-26)$$

Curves $U_p = f(X)$, calculated by formula (3-26), are approximately/exemplarily linear and have different slope/transconductance. Greatest slope/transconductance for the latter/last cascade/stage.

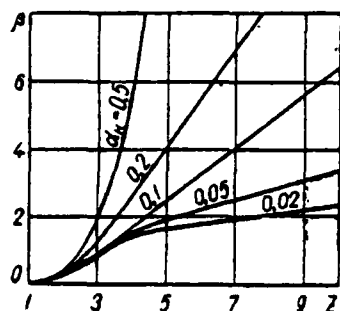


Fig. 53. Generalized curves with the experimental characteristic of UP.

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It is not difficult to see that precise LAKh in this case can be realized with the help of ARU forward in the amplifier, made on the block diagram, analogous to the block diagram, depicted in Fig. 45c. For ARU back, made on the diagram, represented in Fig. 45c, d, LAKh can be realized with the simultaneous, but not identical work of the adjustable cascades/stages. The calculation of the modes of operation of such cascades/stages is very complex.

Tentatively diagram of ARU, made on the diagram for Fig. 45c, can be designed as follows. First they calculate from formula (3-26) of dependence $U_{\text{pr}} = f(Z)$, which have exponential character. Curves $U_{\text{pr}} = f(Z)$ approximate by inclined straight lines, on which the required transmission factors K_{pr} of regulators are determined. The

final fitting of amplifier with LAKh is realized experimentally by a change in the transmission factor K_{pm} of cutoff voltages and initial modes of operation of UK.

The obtained relationships/ratios can be used during the amplification of pulse and continuous harmonic signals for calculating the amplifiers of the bases of self-bias. During the amplification only of pulse radio signals it is possible to calculate amplifiers, also, with the self-bias.

During the calculation of the controlling voltage/stress in amplifiers with the self-bias, intended for amplifying the continuous harmonic oscillations, it is necessary to consider the reaction of the circuit of self-bias (in the vacuum-tube amplifiers) or circuit of heat stabilization (in the transistor amplifiers). Character and degree of reaction depends on the form of passage characteristic of UP and value of the reinforced signal. With the increase of signal from some specific signal level, ARU due to the nonlinearity of passage characteristic of UP comes into action. As a result is observed the combined gain control of cascade/stage both due to U_p , that coming from regulator and due to a change in the voltage/stress of self-bias, caused by a change in the constant component of anode (collector) current.

Functional amplifiers are, based on the nonlinearity of the characteristics of amplifier instruments.

Principle of the operation of diagram.

The automatic gain control due to the nonlinearity of UP can be used in the amplifiers of harmonic oscillations both on the tubes and on the transistors.

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The simplified circuits of cascades/stages with ARU, based on the nonlinearity of UP, are depicted in Fig. 54. Cascades/stages can be made on any of the amplifier circuits, given in Fig. 54, or on one of the cascode diagrams. However, in all diagrams must be general/common/total element/cell - sufficiently high controlling resistor/resistance R , on which the constant component of voltage/stress ΔU , caused by nonlinearity of UP, is selected.

In the case of transistor amplifiers the controlling resistor/resistance in principle can be connected with the emitter circuit or base, but, for representing an improvement in the temperature operational stability, it should be switched on only in the emitter circuit in any circuit diagram of transistor.

The diagram, depicted in Fig. 54c, differs from diagrams by 54a, b and d only in terms of the polarities of the voltages/stresses, applied to the electrodes of transistor, and to opposite directions of the direct currents, flowing on the circuits of diagram.

In the external outline the diagrams in question do not differ from the usual diagrams of selective cascades/stages with the self-bias (or temperature compensation for the transistors). Difference consists of the values of elements/cells R and C of the circuit of self-bias.

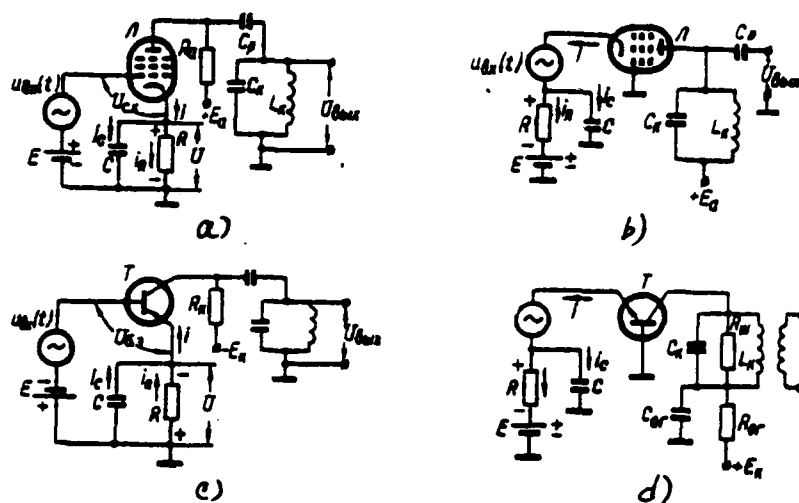


Fig. 54. Diagrams of resonance cascades/stages with ARU, based on the nonlinearity of the characteristics of amplifier instruments with common: a) cathode, b) grid; c) emitter; d) base.

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The high resistor/resistance R , connected with the circuit of self-bias, is shunted by capacity/capacitance C , whose value is selected so that its resistor/resistance for the reinforced high-frequency signals is small, and for the signal amplitude envelope (by slowly varying component of signal current) - very large. In the absence of signal on the input of the cascade/stage through resistor/resistance of R the direct current of amplifier instrument (cathode or emitter) flows/occurs/lasts and on it is selected, large direct/constant voltage U , which by negative

(diagrams a, b, c in Fig. 54) or positive (diagram c in Fig. 54) potential is applied to glass of tube or to the base of transistor. For compensation U and creation of normal bias voltage U_{cm} ($U_{a.n}$ or $U_{a.o}$) into the circuit of the grid of tube (or the base of transistor) the compensating source of positive (negative) voltage E with the low internal resistor/resistance is connected.

The low internal resistor/resistance of source E is necessary for the exception/elimination of effect on the work of the system of ARU of grid (on the high signal levels) or base current, and also for increasing the operational stability of transistor diagram with a change in the ambient temperature. The value of voltage/stress E for the transistor amplifiers is chosen from the condition

$$|\pm E| - |\mp U_n| = |\pm U_{a.n}|, \quad (3-27)$$

for the electron-tube ones -

$$| -U_n | - | +E | = | -U_{a.n} |, \quad (3-28)$$

where U_n - initial voltage/stress on resistor/resistance of R in the circuit of self-bias with absence of signal; $U_{a.n}$ - initial bias voltage.

For the transistor voltage amplifiers $U_{a.n}$ is chosen by the usual method, on the basis of the specifications of transistor, given

in the manual.

For the vacuum-tube amplifiers the operating point is chosen not on the middle of the linear part of the passage characteristic at the point of the greatest slope/transconductance, as is done in usual UK, but on the lower part of the linear section of characteristic. In this case the control of the amplification of cascade/stage begins with comparatively small input voltage and at the same time is preserved the sufficiently large amplification of low signals.

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The automatic gain control in the diagram occurs as follows. During the amplification of low signals (for transistor amplifiers $U_{in} \leq 10$ mV) the amplifier works in the linear conditions, since the nonlinearity of characteristic of UP is not manifested and the constant component of cathode (emitter) current is not changed. With an increase in the amplitude of input signal the nonlinearity of characteristic of UP is manifested and the constant component of cathode $I_{k.0}$ (emitter $I_{e.0}$) current increases. Direct/constant

voltage U on resistor/resistance of R and bias voltage in the vacuum-tube amplifier (Fig. 48) increases, and, on the contrary, is reduced bias voltage in the transistor amplifier, operating point A is displaced to the left, average/mean mutual conductance and transmission factor of cascade/stage are reduced. With a considerable increase in amplitude U_m of input signal the cascade/stage passes into the operating mode with the cutoff of cathode (emitter) current.

During the amplification of radio pulses on resistor/resistance of R is developed a voltage of the detected video pulse (pulse envelope), which is the voltage/stress of ARU. The speed of gain control depends on values of R and C . It is obvious that for the realization of ARU in the pulse amplifier the time of the charge of capacity/capacitance C must be considerably less than the duration of pulse t_n .

The effect of the control of amplification is determined by value R , slope/transconductance and curvature of passage characteristic of UP. Initial slope/transconductance, in turn, depends on value $U_{0.1}$. The greater the resistor/resistance, the greater the effect of control.

Thus, in the diagram in question are combined the functions of the amplification of the high-frequency oscillations and detection.

Because of this there is no need for applying special detector and dc amplifier in the circuit of gain control. In the diagram is realized local ARU, whose operating speed in UK, made on the tube or the high-frequency transistor, can be obtained sufficiently high.

Analysis of the work of amplifier.

As a result of analysis it is necessary to define how the transmission factor of cascade/stage with a change of the input signal in ShDD is changed, i.e., what form of the amplitude characteristic of cascade/stage depending on the initial position of operating point (from the value of voltage/stress $U_{c.u}$) and value of resistor/resistance R. For this it is necessary to define how the constant component of cathode (emitter) current and the fundamental harmonic of anode (collector) current in the dependence on the amplitude of input signal is changed.

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Let us analyze for two most widely used approximations of passage characteristic of UP: quadratic (for the amplifier on the tube with the common cathode) and exponential (for the transistorized amplifier with the general/common/total emitter). In this case we will consider that UP are inertia-free. Subsequently let us take into

account their inertness during the analysis of transistor amplifier.

Quadratic approximation of passage characteristic of UP. As shown in works [8, 11, 12], by the polynomial of the second power in the defined section it is possible to approximate the characteristic of both the tube and the transistor. For the tube this approximation is suitable in the larger measure. Therefore let us analyze in connection with amplifier on the tube, connected with the common cathode.

However, the obtained conclusions/outputs will be accurate both for the transistorized amplifier of type n-p-n and also for the transistorized amplifier of type p-n-p only taking into account the signs of voltages/stresses in accordance with formula (3-27).

Total voltage/stress, which functions at the input of cascade/stage (Fig. 54a),

$$e_1 = u_{bx}(t) + E - U. \quad (3-29)$$

With input signal $u_{bx}(t) = U_m$ for the cathode current

$$I_k = [a_0 + a_1(E - U) + a_2(E - U)^2 + 0,5a_3U_m^2] + [a_1 + 2a_2(E - U)]U_m \cos \omega t + 0,5a_3U_m^2 \cos 2\omega t. \quad (3-30)$$

Constant component (more accurately, slowly changing) $I_{k.0}$ the

amplitude of the first I_{n1} and the second I_{n2} of the harmonics of the cathode current

$$I_{n,n} = a_0 + a_1(E - U) + a_2(E - U) + 0,5a_3U_m^2; \quad (3-31)$$

$$I_{n1} = [a_1 + 2a_2(E - U)] U_m; \quad (3-32)$$

$$I_{n2} = 0,5a_3U_m. \quad (3-33)$$

The first and second harmonics without difficulty pass through capacity/capacitance C, without creating on it noticeable voltages/stresses. The constant component, passing through nucleus CR, creates on it the voltage/stress, which increases in the value in proportion to the charge of capacity/capacitance C. Charge rate C determines operating speed of ARU.

Conservative value of voltage/stress U

$$U = I_{n,n}R. \quad (3-34)$$

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Expressions (3-31) and (3-32) are transcendental and cannot be solved analytically relatively $I_{n,n}$ and I_{n1} . For determination I_{n1} at the given value of the amplitude of input voltage U_m let us compose the equation of Kirchhoff for the cathode circuit (Fig. 54a) and will solve him relative to conservative value of voltage/stress

U:

$$I_{\Sigma}(t) = I_R(t) + I_C(t), \quad (3-35)$$

where $I_R(t)$ - slowly changing current through resistor/resistance of R; $I_C(t)$ - slowly changing current through capacity/capacitance C.

Since

$$I_R(t) = \frac{U(t)}{R} \text{ и } I_C(t) = C \frac{dU}{dt},$$

taking into account expressions (3-31) and (3-35), we obtain

$$C \frac{dU}{dt} + \frac{U}{R} = a_0 + a_1(E - U) + a_2(E - U)^2 + 0,5a_3U_m^2.$$

After introducing designation $E - U = U_{cm} = x$ and taking into account that

$$\frac{dU}{dt} = -\frac{dx}{dt},$$

finally we obtain

$$C \frac{dx}{dt} = a_2x^2 + \left(a_1 + \frac{1}{R}\right)x + \left(a_0 - \frac{E}{R} + 0,5a_3U_m^2\right). \quad (3-36)$$

In steady-state mode/conditions $x = x_y$ and $\frac{dx_y}{dt} = 0$. Therefore

$$a_2x_y^2 + \left(a_1 + \frac{1}{R}\right)x_y + \left(a_0 - \frac{E}{R} + 0,5a_3U_m^2\right) = 0.$$

Whence

$$x_y = U_{cm, y} = \frac{-\left(a_1 + \frac{1}{R}\right) + \sqrt{\left(a_1 + \frac{1}{R}\right)^2 - 4a_2\left(a_0 - \frac{E}{R} + 0,5a_3U_m^2\right)}}{2a_2}. \quad (3-37)$$

In the absence of the voltage of signal $U_m = 0$ and $x = x_u = U_{c.u}$

$$U_{c.u} = x_u = \frac{-(a_1 + \frac{1}{R}) + \sqrt{(a_1 + \frac{1}{R})^2 - 4a_2(a_0 - \frac{E}{R})}}{2a_2}. \quad (3-38)$$

Thus, voltage/stress U_{cm} in the control varies from $U_{c.u}$ to $U_{c.r}$. After substituting expression (3-37) in (3-32), we will obtain formula for calculating the value of the fundamental harmonic of cathode current with the given amplitude of the input signal

$$I_{m1} = U_m \left[-\frac{1}{R} + \sqrt{(a_1 + \frac{1}{R})^2 - 4a_2(a_0 - \frac{E}{R}) - 2a_2^2 U_m} \right]. \quad (3-39)$$

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It is necessary to note that from formula (3-39) it is possible to calculate amplifiers for comparatively low input signals

($U_{m \max} < U_0 - U_{c.u} \approx 2e$ - for the vacuum-tube amplifier;

$U_{m \max} < U_{c.u} - U_0 \approx 0.2e$ - for the transistor). With high levels of the signal, when UP works with the current cutoff, it is expedient to approximate passage characteristic by the broken straight line (Fig. 48) with $p=1$.

Let us find the connection/communication between current $I_{u.u}$ and angle of cutoff θ , for which let us register the equation of the cathode characteristic:

$$i_u = S_u(U_0 + U_u) = S_u(U_0 + E - I_{u.u}R + U_m \cos \omega t), \quad (3-40)$$

where $S_k = \text{tg} \alpha$ - slope/transconductance of cathode characteristic.

The dc current component $I_{k, \text{av}}$ during the resolution of current i_k in Fourier series

$$I_{k, \text{av}} = \frac{S_k [\theta (U_0 + E) + U_m \sin \theta]}{\pi + S_k R \theta}. \quad (3-41)$$

After taking in formula (3-40) $\omega t = 0$, we will obtain

$$I_{k, \text{av}} = \frac{U_0 + E + U_m \cos \theta}{R}. \quad (3-42)$$

During the joint solution of equations (3-41) and (3-42)

$$U_m = \frac{\pi (U_0 + E)}{S_k R \sin \theta - (\pi + S_k R \theta) \cos \theta}. \quad (3-43)$$

For the transistor of type p-n-p (according to Fig. 48b)

$$I_b = S_b (-U_0 + e_1); \quad (3-44)$$

$$I_{b, \text{av}} = \frac{-U_0 + E + U_m \cos \theta}{R}; \quad (3-45)$$

$$U_m = \frac{\pi (E - U_0)}{S_b R \sin \theta - (\pi + S_b R \theta) \cos \theta}. \quad (3-46)$$

Dependences $I_{k, \text{av}} = f(U_{m, \text{av}})$ and $I_{k1} = f(U_{m, \text{av}})$ for the tube on the high signal level are calculated in the following order. First by formula (3-43) is found dependence $U_m = f(\theta)$, being assigned by values θ , and then according to formula (3-41) or (3-42) for given values U_m and θ is calculated conductivity $I_{k, \text{av}} = f(\theta)$.

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After this from tabular values of the coefficients of expansion $\alpha_0(\theta)$ by constant component and $\alpha_1(\theta)$ of the fundamental harmonic of anode current are calculated dependence $I_{a1} = f(U_m)$, and then amplitude characteristic of cascade/stage according to the formula

$$U_{\text{amx}} = I_{a1} R_0, \quad (3-47)$$

where $I_{a1} = I_{a1} - I_{a1} = (0.75 \div 0.8) I_{a1}$ - amplitude of the fundamental harmonic of anode current; R_0 - the total resistance of load.

For the transistors the calculation is carried out analogously.

During the calculation of AKh of the cascade/stage, loaded to the analogous adjustable cascade/stage, one should remember that its load is shunted by the input resistance of the following cascade/stage, which is changed with an increase in the signal.

Calculated curves of the dependence of the bias voltage and fundamental harmonic of anode current for the tube 6Zh3P at two values of initial bias voltage $U_{0.m}$ and different values of resistor/resistance of R are depicted in Fig. 55. The same figure shows the experimental points, which coincide sufficiently well with the calculated ones.

Being congruent/equating Fig. 55b with Fig. 13, we see that curves $I_{n1} = I(U_{n1})$ by nature coincide sufficiently well with the curves $z=f(x)$ for the logarithmic amplifier. Therefore it is possible to carry out a successive work of cascades/stages and to obtain precise LAKh of multistage amplifier in ShDD.

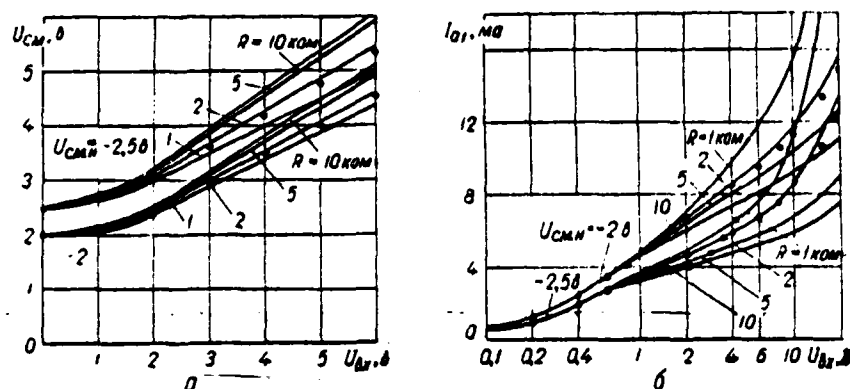


Fig. 55. The curves of the change: a) bias voltage; b) the fundamental harmonic of the anode current of tube 6Zh3P in the dependence on the amplitude of input voltage ($U_{in} = 220$ c; $U_{in} = 150$ c).

Key: (1). kilohm.

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Being assigned a different value of the resistors/resistances of plate load, from curves $I_{a1} = f(U_{in})$ several sets of amplitude characteristics of cascade/stage calculate, choosing that characteristic, which satisfies the successive work of cascades/stages. However, it is necessary to note that with the large input signals the characteristic of cascade/stage sharply is bent due to by-passing of the following cascade/stage, in consequence of which the extent of the quasi-linear section of characteristic is limited.

Therefore it is not possible to carry out a successive work of more than three (maximum of four) cascades/stages and to obtain precise LAKh in the range of more than 50-60 dB. The schematic of logarithmic amplifier is given in Fig. 56. Parameters of amplifier at the work in the linear conditions: $f_0=30$ MHz; $K_0=10^3$; $\Pi=1.2$ MHz with the work in the logarithmic mode/conditions: $D=60$ dB; the slope/transconductance of LAKh $\sigma=0.3$ V/dB; accuracy of LAKh $\delta U \approx 3\%$.

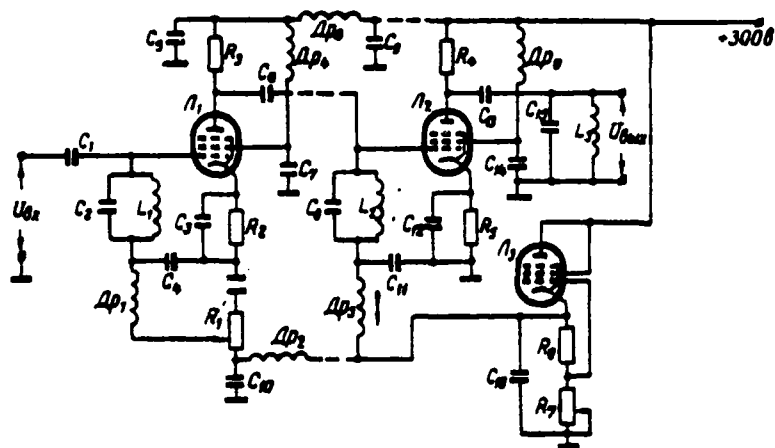


Fig. 56. Basic schematic of logarithmic amplifier with ARU on pulse envelope: L_1, L_2 - 6Zh1P; R_2, R_3, R_4, R_5 - 2 kilohm; R_1 - 10 kilohm; R_6 - 150 ohm; R_7 - 3.3 kilohm; C_1, C_2, C_3 - 300 pF; C_4, C_{12} - 180 pF; $C_5, C_6, C_7, C_8, \dots$ - 3200 pF.

Key: (1). In.

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The advantages of the schematic of logarithmic amplifier with ARU due to the nonlinearity of the characteristic of the tube:

1) the high stability of the parameters of amplifier, which is ensured by a deep negative feedback (OOS) in the direct current and by the absence of nonlinear semiconductors;

2) the short time of gain control, which does not exceed the tenths of microsecond;

3) the small overall sizes of amplifier, equal to the overall sizes of usual selective amplifier.

With an increase of resistor/resistance of R in the cathode circuit the bend of amplitude characteristic increases, in this case the action of the destabilizing factors sharply descends. With the scatter of the parameters of tubes to 30% factor of amplification of cascade/stage in the linear conditions remain in effect constant, and the form of amplitude characteristic is changed insignificantly. Because of this it is possible to fulfill multistage amplifier-limiter with the sufficiently rigid and stable limitation in ShDD under the influence of different destabilizing factors (change in the anode voltage, filament voltage, temperature, etc.). In particular, the logarithmic amplifier with the addition of the detected voltages/stresses, whose schematic diagram is given in Fig. 74, on the output of detector is amplifier-limiter.

Approximation of the passage characteristic of amplifier instrument by exponential curve. Since exponent to the greatest

extent answers the passage characteristic of the transistors, let us consider diagram of ARU in the transistorized amplifier, connected on the common-emitter connection,. The controlling resistor/resistance R is connected with the emitter circuit.

Then for the emitter characteristic

$$I_e = A_e e^{\gamma(E-U)} e^{\gamma U_m \cos \omega t}. \quad (3-48)$$

Constant component and fundamental harmonic of emitter current according to expressions (2-76) and (2-77)

$$I_{e,0} = A_e e^{\gamma U_{e,0}} I_0(\gamma U_m); \quad (3-49)$$

$$I_{e,1} = 2A_e e^{\gamma U_{e,0}} I_1(\gamma U_m), \quad (3-50)$$

where $I_0(\gamma U_m) = J_0(j\gamma U_m)$ and $I_1(\gamma U_m) = -jJ_1(j\gamma U_m)$ - moduli/modules of complex variable functions.

Since

$$U = I_{e,0} R \quad (3-51)$$

and

$$U_{e,0} = E - U, \quad (3-52)$$

it is possible to register

$$U = A_e R e^{-\gamma U} I_0(\gamma U_m). \quad (3-53)$$

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Expression (3-52) is transcendental. Approximate analytical solutions, and also methodology of the graphical solution of this equation are given in work [8]. Approximate solutions relative to voltage/stress have the following expressions:

$$U = \frac{-[1.1\gamma - \gamma E + 2.2 - k] + \sqrt{[1.1\gamma - \gamma E + 2.2 - k]^2 + 4.4\gamma[\gamma E + \ln k + 2]}}{2}, \quad (3-54)$$

where

$$k = \ln A_0 R I_0 (\gamma U_m);$$

$$U = \frac{\gamma E + 0.68 + \ln A_0 R I_0 (\gamma U_m)}{\gamma + 0.8}. \quad (3-55)$$

For $E > U_{0.6}$, which very frequently is performed, equation (3-52) relatively $U_{0.6}$ to solve is comparatively simple. After taking in expression (3-52) $|E| > |U_{0.6}|$, it is disregarded by value $U_{0.6}$ in comparison with E we take the logarithm of. As a result

$$U_{0.6} = \frac{1}{\gamma} \ln \frac{E}{A_0 R I_0 (\gamma U_m)}. \quad (3-56)$$

or

$$U = E - \frac{1}{\gamma} \ln \frac{E}{A_0 R I_0 (\gamma U_m)}. \quad (3-57)$$

As a result of the calculations conducted it is

established/installed, that the relative disagreement in the solutions of equation (3-52) in calculator and for approximation formula (3-54) does not exceed 1%, and during the calculation according to approximate formula (3-57) this disagreement does not exceed 5%. Therefore for the estimate calculations it is possible to apply formula (3-57).

After the calculation of dependence $U_{\text{с.с}} = f(U_m)$ according to formula (3-56) taking into account expressions (3-52) and (3-50) is calculated the dependence of the amplitude of the fundamental harmonic of collector current $I_{\text{с.с}} = f(U_m)$ on the amplitude of input signal, and then from formula (3-47) the AKh of cascade/stage.

The given formulas can be used for calculating the amplitude characteristic at the sufficiently low frequencies for this type of transistor $\omega < \omega_{\text{rp}}$, when the inertness of the transistor can be disregarded/neglected. Then the coincidence of experimental data with the calculated is obtained very good [13].

However, at frequencies $\omega > \omega_{\text{rp}}$ it is necessary to consider the inertial properties of transistor.

Let us consider one of the possible versions of the calculation of AKh of cascade/stage taking into account the frequency dependence of the conductivity of direct drive Y_{21} .

At the low frequency the conductivity of the direct drive of transistor according to the collector characteristic

$$Y_{21 \text{ a. q.}} = \left. \frac{\partial I_K}{\partial U_{\text{a. o.}}} \right|_{U_{\text{K. o.}} = \text{const}} = A_K \gamma e^{U_{\text{a. o.}}} \quad (3-58)$$

For conductivity Y_{21} is known the following dependence on the frequency:

$$\dot{Y}_{21} = \frac{Y_{21 \text{ a. q.}}}{1 + j \frac{\omega}{\omega_{\text{rp}}}} = Y_{21 \text{ a. q.}} \dot{m},$$

where $\dot{m} = \frac{1}{1 + j \frac{\omega}{\omega_{\text{rp}}}}$ - factor, which calculates the decrease of conductivity Y_{21} on the basis of frequency; ω - frequency of the amplified oscillations; ω_{rp} - upper cut-off frequency for this diagram.

For the operating frequency ω expression (3-58) takes the form

$$\left. \frac{\partial I_K}{\partial U_{\text{a. o.}}} \right|_{U_{\text{K. o.}} = \text{const}} = \dot{m} A_K \gamma e^{U_{\text{a. o.}}} \quad (3-59)$$

If we disregard/neglect loading effect (cascade/stage is loaded to a comparatively low input resistance of the following

cascade/stage), then from formula (3-59)

$$\frac{dI_n}{dU_{a.o}} = mA_n \gamma e^{U_{a.o}} \approx |Y_{21}| e^{\gamma}. \quad (3-60)$$

Since during the construction of logarithmic amplifiers it is important to know amplitude relationships/ratios, phase factor $e^{j\gamma}$ it is possible not to consider. Certain error, which is obtained in this case and caused by the dependence of phase on $U_{a.o}$, can be reduced during the adjustment of amplifier. Analysis and calculations with this assumption considerably are simplified. Expression (3-60) can be now registered in the form

$$\left| \frac{dI_n}{dU_{a.o}} \right| = mA_n \gamma e^{U_{a.o}},$$

where m - modulus/module of coefficient $m(\omega)$, determined by expression (2-95).

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After the integration

$$I_n = mA_n e^{U_{a.o}} + C. \quad (3-61)$$

We find integration constant C , after taking the current of collector/receptacle equal to $I_{n.c.n.}$ with the voltage/stress between the emitter and the base, equal to $U_{a.o.n.}$. Substituting the value of C into expression (3-61), we obtain

$$I_n = mA_n (e^{U_{a.o}} - e^{U_{a.o.n.}}) + I_{n.c.n.}$$

Since at the input of cascade/stage the oscillatory circuit is connected, the voltage of signal is harmonic. In this case

$$U_{a,0} = U_{a,0,z} + U_m \cos \omega t.$$

Taking into account the expansion of expression (2-74) we determine constant component and fundamental harmonic of the current of emitter (collector/receptacle):

$$I_{a,z} = I_{a,z,z} + A_m m e^{U_{a,0,z}} I_0(\gamma U_m) - A_m m e^{U_{a,0,z}}; \quad (3-62)$$

$$I_{a1} = 2A_m m e^{U_{a,0,z}} I_1(\gamma U_m). \quad (3-63)$$

Direct/constant voltage between the base and the emitter

$$U_{a,0,z} = U_{a,0,z} - \Delta U_{a,0,z},$$

where $\Delta U_{a,0,z}$ - incremental stress $U_{a,0,z}$, caused by current $I_{a,z}$.

As we see, expressions for I_{a1} and $I_{a,z} \approx (I_{a,z})$ can be registered

so:

$$I_{a,z} = I_{a,z,z}(1 - m) + m I_{a,z,z} e^{-\gamma U_{a,0,z}} I_0(\gamma U_m); \quad (3-64)$$

$$I_{a1} = 2m I_{a,z,z} e^{-\gamma U_{a,0,z}} I_1(\gamma U_m). \quad (3-65)$$

After simple conversions from expression (3-64) we obtain

$$I_0(\gamma U_m) = \left(1 + \frac{\Delta U_{a,0,z}}{m R I_{a,z,z}}\right) e^{\gamma U_{a,0,z}} \quad (3-66)$$

From expressions (3-64), (3-65), (3-66) it is possible to comparatively easily calculate the amplitude characteristic of cascade/stage. For this should be assigned different values $\Delta U_{a. a.}$ at the given values of resistor/resistance of R and current of emitter $I_{e.}$ determined values $I_0(\gamma U_m)$, and then on tables $I_0(\gamma U_m)$ and $I_1(\gamma U_m)$ determined values $x = \gamma U_m$, U_m and $I_1(\gamma U_m)$. After substituting obtained values $I_1(\gamma U_m)$ into expression (3-66), let us find the amplitude of the fundamental harmonic of collector current, and from it and the resistance/resistor of load - amplitude of output voltage/stress.

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During comparatively high resistors/resistances of R, when inequality $\Delta U_{a. a.} \ll m R I_{e.}$ is fulfilled i.e. When the voltage of the source of the bias/displacement is of much more than initial bias/displacement on the base transistor, calculations considerably are simplified. Expression (3-66) takes the form

$$I_0(\gamma U_m) \approx e^{\gamma \Delta U_{a. a.}}$$

After its substitution into expression (3-65) we obtain

$$I_{m1} = 2m I_{e.} \frac{I_1(\gamma U_m)}{I_0(\gamma U_m)} \quad (3-67)$$

For the output voltage/stress of the cascade/stage

$$U_{\text{max}} = \frac{I_{\text{K1}}}{Y_0} \quad (3-68)$$

where Y_0 - general/common/total load admittance at the resonance frequency.

With the work of cascade/stage in the linear conditions argument γU_m of the Bessel functions is small. In this case it is possible to register:

$$\begin{aligned} I_1(\gamma U_m) &\approx \frac{\gamma U_m}{2}; \\ I_0(\gamma U_m) &\approx 1. \end{aligned}$$

Then the transmission factor of cascade/stage in the linear conditions

$$K_n = \frac{m \gamma I_{\text{K. n. n}}}{Y_0} = \frac{m S_n}{Y_0}. \quad (3-69)$$

Amplitude is the characteristic of cascade/stage with the work in the nonlinear mode/conditions

$$U_{\text{max}} = 2 \frac{K_n}{\gamma} \cdot \frac{I_1(\gamma U_{\text{max}})}{I_0(\gamma U_{\text{max}})}. \quad (3-70)$$

It is very simple to calculate the amplitude characteristic of cascade/stage from formula (3-70). For this it is necessary to know the transmission factor of cascade/stage with the work in the linear conditions, coefficient γ and to have appropriate tables $I_1(\gamma U_m)$ and $I_0(\gamma U_m)$.

Characteristics $I_{out} = f(U_m)$ for the different values of resistor/resistance of R are depicted in Fig. 57. During the calculation it is accepted $\gamma=15$. From expression (3-70) and Fig. 57 it is evident that during the high resistors/resistances of R , when inequality $U_{a.e.} \ll mKI_{a.e.}$ is fulfilled amplitude characteristic virtually does not depend on value R and has form requiring, for obtaining LAKh by the method of adding, the voltages/stresses from the outputs of amplifier stages.

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During completely specific resistor/resistance of R it is possible to obtain the virtually linear characteristic of cascade/stage in the considerable dynamic range. This fact can be used for the linearization of the passage characteristic of transistor in obtaining FAKh with the help of the nonlinear elements/cells, included in the load circuit or OOS.

At the high values of argument γU_m functions $I_0(\gamma U_m)$ and $I_1(\gamma U_m)$ become approximately/exemplarily equal and, according to expression (3-67), the amplitude of the fundamental harmonic of collector current becomes constant (broken line in Fig. 58a). Thus, constancy I_{m1} and, consequently, output voltage/stress, can be observed even when $U_{m \max} < U_n$ and the limitation on the collector circuit, which usually is observed upon reaching/achievement of alternating voltage on the load of the supply voltage, did not appear. However, as can be seen from experimental curves, the output voltage/stress of cascade/stage grows with an increase in the input voltage. This is explained by the special feature/peculiarity of the work of transistor on the high signal level, which consists in the fact that with the considerable voltages/stresses between the base and the emitter transition the base - collector/receptacle in the diagram with OE or transition emitter - collector/receptacle in the diagram OB works as usual diode.

As a result it is obtained, that the amplitude characteristic on its form does not satisfy the requirements both to the successive work of nonlinear cascades/stages and to the method of adding the output voltages/stresses.

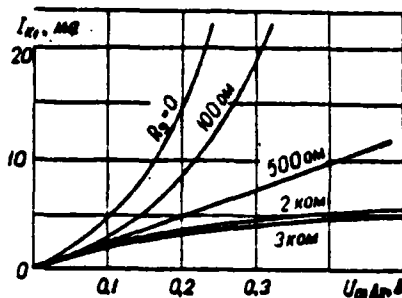


Fig. 57. Curves of the dependence of the fundamental harmonic of the current of collector/receptacle in the transistor amplifier with ARU due to the nonlinearity of the characteristic of transistor.

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During precise calculations of AKh it is necessary to consider the effect of the input resistance of the following cascade/stage, whose value for resonance turn-on frequency of transistor on the diagram with OB and with OE can be designed respectively from the formulas

$$R_{\text{ex. o. s}} = \frac{U_{\text{m ex}}}{I_{\text{e1}}}; \quad (3-71)$$

$$R_{\text{ex. o. s}} = \frac{U_{\text{m ex}}}{I_{\text{b1}}}, \quad (3-72)$$

where I_{e1} and I_{b1} - amplitude of the fundamental harmonic of respectively emitter and base current.

Currents I_{a1} and I_{c1} can be calculated from formula (3-67), after replacing in the formula current $I_{k, n, n}$ and coefficient γ by currents $I_{a, n, n}$ or $I_{c, n, n}$ and γ_a or γ_c where γ_a and γ_c - coefficients of the approximation of characteristics $i_a = f(U_{a, a})$ and $i_c = f(U_{c, c})$.

It is possible to show that with an increase in input voltage U_m resistor/resistance R_{ax} grows also with the work of cascade/stage in the mode/conditions with the small angle of cutoff of collector current (0-90°) it becomes such large that it is possible not to consider in comparison with the resistor/resistance, shunting the oscillatory circuit $R_m = 1-2$ kilohm.

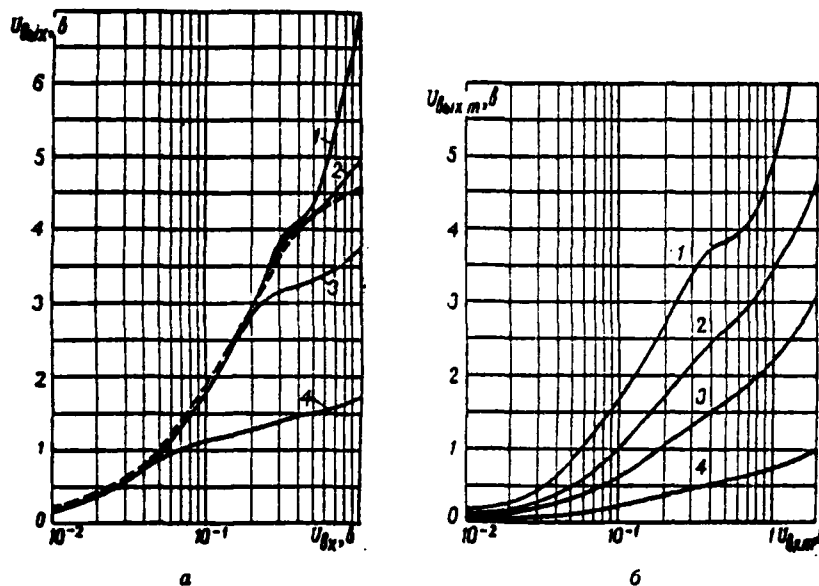


Fig. 58. Amplitude characteristics of cascade/stage with ARU due to the nonlinearity of the characteristic of a transistor of the type P411: a) at the different values of voltage on the collector/receptacle of transistor and $R_c = 1$ kilohm, $R_{\text{feedback}} = 1$ kilohm, $I_{c, \text{stat}} = 4$ mA; b) at the different values of the resistor/resistance of feedback and $R_c = 4.3$ kilohm; ---- calculated; — experimental.

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For calculating of AKh the cascade/stage according to formula (3-70) at the high values of the amplitude of input voltage $U_m > 0.5$ in the dependence on the coefficient γ are required the values of the moduli/modules of functions $I_0(x)$ and $I_1(x)$ for argument $x = \gamma U_m > 10$. The table of the moduli/modules of these functions with a change in the

argument from 10 to 40 is given in the application/appendix of work [8]. From it it is possible to calculate AKh of cascade/stage to the values of input voltage $U_m = 1-3$ v in the dependence on the value of coefficient γ . This it is sufficient for the development/detection of the character of amplitude characteristic. With further increase in amplitude U_m amplitude characteristic linearly grows, in consequence of which there is no need for calculating it from formula (3-70).

Logarithmic amplifier to are and by radio-output.

For obtaining precise LAKh of multistage amplifier with the radio-output nonlinear cascades/stages must have the amplitude characteristics, described by expressions in Tables 1 and 3,. The special feature/peculiarity of these characteristics is the presence of quasi-linear section, i.e., a finite increment in the output voltage/stress with the differential amplification factor, equal to coefficient $a = \frac{1}{\ln N}$.

AKh required for the successive work of cascade/stage with the quasi-linear section can be obtained in the cascade/stage with ARU, based on the nonlinearity of transistor, if we consecutively/serially with the adjustable resistor in the emitter circuit include/connect the resistor/resistance of feedback small in the value.

The schematic diagram of four-stage resonance logarithmic amplifier with the radio-output and its description are given in work [8].

For obtaining LAKh by the method of adding the voltages/stresses from the outputs of amplifier stages in AKh of cascade/stage on the high signal levels must be the horizontal section, whose extent must be not less dynamic range d of the nonlinear (or logarithmic) section of characteristic.

The form of AKh of cascade/stage with further increase of signal does not have vital importance and can be any, since the voltage on the input of this cascade/stage ceases to grow due to the limitation in the previous cascade/stage.

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The simplest and most efficient method of obtaining the horizontal section of characteristic is the inclusion into the collector circuit of effective resistance R_{or} (Fig. 54d). During the amplification of continuous oscillations as resistor/resistance R_{or} ^{it is possible to use} λ the resistor/resistance of filter R_{ϕ} can be used. With a sharp increase of the constant component of collector current descends the supply voltage on the collector/receptacle of transistor and, consequently, also the level of limitation. Usually the value of resistor/resistance R_{or} lies/rests within the limits of 300-500 ohms.

For the efficient action of resistor/resistance R_{or} during the amplification of pulse signals the time constant of chain/network $\tau_{or} = R_{or}C_{or}$ must be considerably less than the duration of the pulse of signal t_{π} .

With fulfilling of inequality $\tau_{or} = (2+3)t_{\pi}$ due to chain/network $C_{or}R_{or}$ is regulated the amplification of cascade/stage on the prolonged interference with unchanged amplification factor in the pulse signal.

ARU ON PENTODE GRID.

The dependence of slope/transconductance on the value of voltage/stress on the pentode $S=f(U_n)$ or heterodyne $S=f(U_r)$ grids for many tubes can be approximated by one or two linear segments with different inclination/slope. During the supplying of controlling voltage/stress U_p on the pentode or heterodyne grid the dependence of the transmission factor of cascade/stage also is linear (straight line ac in Fig. 59) or piecewise-linear (segments ab and bc in Fig. 59)

$$K = \frac{U_{\text{max}}}{U_{\text{ax}}} = K_n - \Delta K(U_p) = K_n \left(1 - \frac{hU_p}{U_n}\right), \quad (3-73)$$

where U_n - cutoff voltage of tube on the pentode grid, with which $K=0$; h - coefficient, which is determining the degree of the effect of the controlling voltage/stress on one cascade/stage.

With strictly linear dependence $S=f(U_n)$ coefficient $h=1$. For segment ab in Fig. 59 coefficient $h_1 = \frac{\lg a}{\lg a_1} < 1$, for bc - $h_2 = \frac{\lg a}{\lg a_2} > 1$.

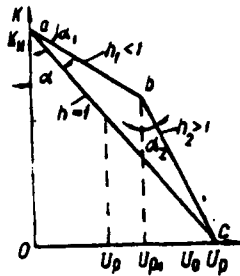


Fig. 59. Curves of dependence $K=f(U_p)$ during the adjustment on the pentode or heterodyne grid of tube.

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The attempt to give the procedure of calculation of multistage amplifier with the adjustment according to the pentode grid of tube is done in work [40]. However, the methodology, given in this work, is fairly complicated and from it it is not possible to accurately calculate the controlling characteristic. The attempt to give simpler and more precise calculation procedure is done below.

Equalizing the right sides of expressions (2-173) and (3-73), we obtain

$$1 - \frac{U_p}{U_0} = \frac{1}{2},$$

whence

$$U_p = U_0 \left(1 - \frac{1}{2}\right). \quad (3-74)$$

Expression

^(3-74) it is useful for calculating self-regulation of cascades/stages $U_p = f(z \text{ and } x)$ multistage FU, made on the block diagram (Fig. 45a) (ARU back or forward).

During the simultaneous adjustment of all cascades/stages n-cascade FU, made on block diagram (Fig. 45c and d), for voltage/stress U_p

$$U_{pt} = U_{ot} \left(1 - \sqrt{\frac{z}{2}} \right). \quad (3-75)$$

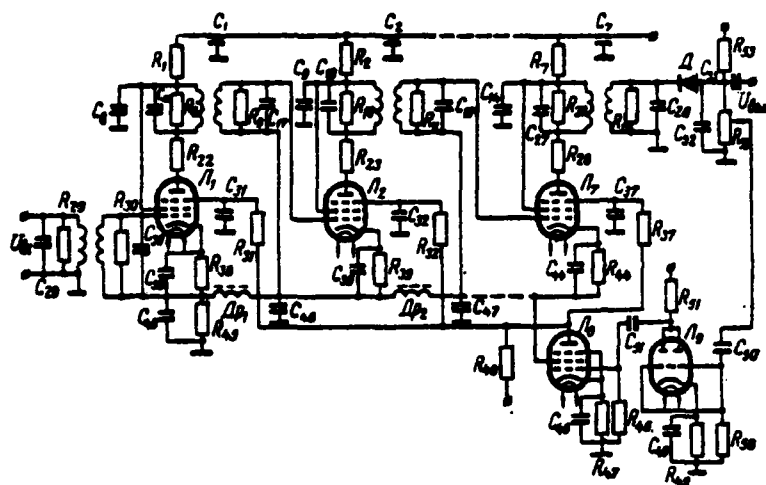


Fig. 60. Schematic diagram of logarithmic amplifier with ARU on the pentode grids of tubes.

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As a result of the conducted calculated and experimental investigations it is established/installed, that the logarithmic amplifier can be performed on any of three diagrams (Fig. 45a, c, d). However, the diagram, given in Fig. 45d, for which relationship/ratio (3-75) is correct is simplest and easily realizable. The curves, calculated by formula (3-75), in their form are analogous to the curves, given in Fig. 51. An example of the calculation of the schematic of the resonance logarithmic amplifier of the pulse radio signals and ARU according to the pentode network/grid, depicted Fig. 60, is given in Chapter 6. Amplifier has following data:

$$K_0 = 10^3; \Pi = 4 \text{ Msz}; f_0 = 30 \text{ Msz}; D_{\text{MAX}} \approx 80 \text{ dB}; \\ U_{\text{EX. H}} = 5 \text{ V}; \sigma \approx 50 \text{ mV/dB}; \delta U \approx 3\%; \delta \sigma = 15\%.$$

Key: (1). MHz. (2). dB. (3). μV . (4). mV/dB.

COMBINED ARU.

Combined ARU by a change in the mode of operation of UP in the direct current can be realized by the simultaneous adjustment:

- 1) with the help of controlling voltage/stress U_p , entering control electrode of UP, together with ARU on the emitter circuit (cathode) due to the nonlinearity of characteristic UP;
- 2) with the help of U_p , that enters the pentode grid of tube, together with ARU through the screen grid due to the increase of the current of the screen grid of tube;
- 3) through the pentode grid with the help of U_p , together with ARU through the cathode circuit due to the nonlinearity of cathode characteristic;
- 4) through the pentode grid with the help of U_p , together with ARU through the screen grid and the cathode circuit.

Let us consider briefly each of the versions.

1. Even if cascade/stage is circuit of self-bias or temperature stabilization, then during the supplying of controlling voltage U_p on control electrode of UP it is necessary to consider effect of these circuits on effect of adjustment. With the linear passage characteristic of UP these circuits reduce the effect of adjustment and, on the contrary, with the nonlinear characteristic of UP on the high signal level these circuits increase the effect of adjustment. As a result combined ARU of the 1st version is obtained.

The schematic diagram of transistor cascade/stage with the simultaneous adjustment due to U_p , developed by regulator (T_1), and resistors/resistances in the emitter circuit R_e is given in Fig. 61.

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For this diagram

$$U_{\text{out}} = U_{\text{a.e.}} = E_p - U_e$$

where

$$E_p = E - U_p; \quad U_e = I_{\text{a.e.}} R_e$$

With the absence of the voltage of signal $U_p = 0$. Then

$$U_{\text{a.e.}} = U_{\text{a.e.}} = E - I_{\text{a.e.}} R_e$$

In order to obtain completely specific AKh of cascade/stage, it is necessary at this value of the amplitude of input signal U_m the initial bias voltage $U_{a. n}$ to decrease by value U_p (Fig. 48b), whose value can be designed on one of formulas (3-14), (3-15), (3-18), (3-19), (3-24), (3-26). With decrease $U_{a. n}$ operating point A on the characteristic is displaced into point A_1 . Due to this, on one hand, are reduced the constant component of emitter current by value $\Delta I'_{a. n}$ and voltage/stress U on resistor/resistance R_e . Operating point attempts to be moved into point A'_1 . On the other hand, due to the nonlinearity the characteristics increase the dc current component of emitter by value $\Delta I''_{a. n}$ and voltage/stress U. Operating point attempts to be shifted into point A_1 . In steady state is performed the following equality:

$$U_{a. n} - U_p = E - U_p' - (I_{a. n. n} - \Delta I'_{a. n} + \Delta I''_{a. n}) R_e$$

or

$$-U_p = -U_p' + (\Delta I'_{a. n} - \Delta I''_{a. n}) R_e$$

Whence value U_p' , which it must manufacture regulator,

$$U_p' = U_p + (\Delta I'_{a. n} - \Delta I''_{a. n}) R_e \quad (3-76)$$

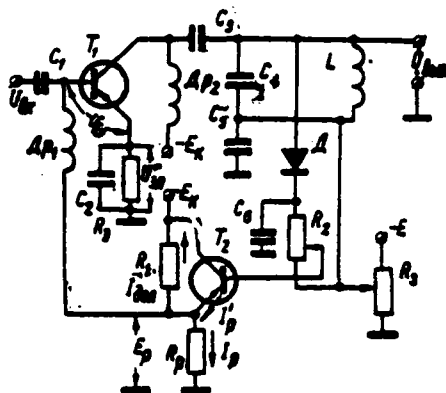


Fig. 61. Schematic diagram of cascade/stage with the simultaneous adjustment.

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Component/term/addend $\Delta I'_{\text{a}}$ in expression (3-76) is caused by the nonlinearity of characteristic of UP, also, on the high levels of the signal, when $U_p \approx U_{c.m.}$ can prove to be congruent with $\Delta I'_{\text{a}}$ or even more than it. In works [34, 35] during the calculation of systems ARU operate with the differential slope/transconductance of passage characteristic and component $\Delta I'_{\text{a}}$ they do not usually consider that, naturally, on the high signal level it leads to the considerable error in calculations.

Dependence $U_p = f(z)$ according to formula (3-76) should be calculated in the following sequence. Dependence $U_p = f(z)$ without

taking into account the effect of controlling resistor/resistance R_s . First is calculated then for given values U_m they determine $\Delta I'_{s, n}$ according to the characteristic of UP and $\Delta I'_s$ - employing procedure and formulas, presented above.

The version of combined ARU examined it is expedient to apply in the amplifiers from the gain control increased by depth with the large signals, in particular, in the amplifier-limiters, in the latter/last cascades/stages of the logarithmic amplifier, made on the block diagram (Fig. 45d).

2. In this case

$$U_p = U'_p + \Delta I_{\text{amp}} R_{\text{amp}} h_{s, n}$$

Whence

$$U'_p = U_p - \Delta U_{\text{amp}} h_{s, n} \quad (3-77)$$

where $h_{s, n} = \frac{S_{p, \text{amp}}}{S_{p, n}}$ - coefficient, which is determining the efficiency of adjustment according to the screen grid relative to the effect of adjustment on the pentode grid; $S_{p, s} = \frac{\Delta I_s}{\Delta U_{\text{amp}}}$ and $S_{p, n} = \frac{\Delta I_n}{\Delta U_n}$ - average/mean slopes/transconductances of self-regulation respectively on the screen and pentode grids.

3. For this version

$$U'_p = U_p - (\Delta I'_n - \Delta I'_s) R_n h_n \quad (3-78)$$

where $h_n = \frac{S_{p, n}}{S_{p, s}}$ and $S_{p, n} = \frac{\Delta I_n}{\Delta U_n}$ - slope/transconductance of

self-regulation on the cathode or, which is the same, on control electrode.

For the majority of tubes an increment in current $\Delta I_k = 0$, since with a change in the voltage/stress on the pentode grid the cathode current remains in effect constant.

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For the realization of the 3rd version during the amplification of continuous oscillations is necessary, that the resistor/resistance in the circuit of screen grid would be absent, i.e., $R_{\text{emp}} = 0$; during the amplification of pulse radio signals necessary, that would be fulfilled the inequalities: $C_p R_{\text{emp}} \gg R_k C_k$; $R_k C_k \ll t_n$.

4. This version is general/common/total. For it

$$U_p' = U_p - \Delta I_{\text{emp}} R_{\text{emp}} h_{\text{emp}} - (\Delta I_k' - \Delta I_k'') R_{11} h_{11} \quad (3-79)$$

§3. Methods of increasing the stability of frequency and phase responses of FU with ARU.

The fundamental reason for the distortion of the frequency and phase responses of amplifier, during the gain control, especially and transistor amplifiers, is a considerable change in active and

reactive components of input and output conductances of UP with a change in the position of operating point. This deficiency/lack to a considerable extent can be reduced, applying:

- 1) the compensating nonlinear elements/cells in the emitter or collector circuits of transistor;
- 2) the controlled attenuators in input or output circuit of the adjustable cascade/stage and in the feedback loop;
- 3) diagram on composite/compound transistors;
- 4) interstage coupling elements with the optimum parameters.

The first three methods of increasing the stability of the characteristics of transistor amplifiers with ARU are quite fully illuminated in the literature [8, 89, 90, 94, 102].

APPLICATION OF COMPENSATING NONLINEAR ELEMENTS.

The essence of this method lies in the fact that for the distortion elimination of the characteristics of the amplifier, parameters of the adjustable transistor caused by change, into the appropriate point of amplifier circuit (to the input or the output)

is switched on the special element/cell, whose parameters into the alignment procedure are changed according to the law, opposite to the law of a change in the parameters of transistor. As a result the partial or complete compensations for changes in the parameters of transistor and the exemplary/approximate constancy of the circuit parameters at the point of the connection of the compensating element/cell is observed.

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For the transistor amplifiers by this compensating element/cell can be any p-n junction, in particular, transistor or semiconductor diode [8, 89, 90, 94, 96, 102], connected with input circuit of cascade/stage during the adjustment a change in the current of emitter or in parallel to load - during the adjustment by a change in the voltage/stress on the collector/receptacle.

Several versions of the inclusion of the compensating elements/cells into input circuit of the adjustable transistor are given in work [8].

The parameters of the compensating elements/cells must be selected so that the total input current of the adjustable cascade/stage in the alignment procedure would remain constant, which

will stipulate the constancy of the input resistance of cascade/stage.

The high stability of the frequency characteristic of multistage amplifier can be obtained only when elements/cells are selected from the condition for the compensation for a change in active and reactive components of input and output conductances of transistors.

If the compensating nonlinear element/cell is connected with the emitter circuit, the virtually constant input resistance of cascade/stage can be obtained in the range of a change in the amplification to 40 dB.

The stability of the frequency characteristic of multistage amplifier is obtained above upon the start of transistors on the diagram with OB. This is explained by the fact that output resistance of transistor with the grounded emitter with one and the same change in the current of emitter. Therefore a change in output conductance more easily to compensate for in the schematic of grounded-base transistor.

The results of the compensation for a change in the output capacitance of the transistor, connected in diagram with OB and OE, are good, it is possible to obtain, switching on the compensating

elements/cells in the circuit of collector/receptacle [8].

APPLICATION OF THE CONTROLLED ATTENUATORS.

With the aid of the electrically controlled attenuators it is possible to change the amplification of cascade/stage during the constant/invariable mode of operation of UP on the direct current. Recently was published a considerable number of works and patents [35, 70, 75, 78, 79, 80, 81, 84, 88, 103, 105, 106], dedicated to a question of the use/application of the controlled attenuators for the realization of ARU in the transistor amplifiers.

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However, in these works the realization of deep ARU, which makes it possible to obtain small changes in the output voltage/stress (in the ideal case, constancy U_{out}) in ShDD of a change in the input voltage examines. The questions, connected with the realization of assigned FAKh with the help of the controlled attenuators, until now, actually are not examined.

The diagrams of the controlled attenuators are given in Fig. 62.

The controlled attenuators (UA) can be connected at the input

either at the output of amplifier stage or in the feedback loop, it is analogous with the starts of nonlinear elements/cells and dividers. For increasing the efficiency of gain control consecutive UA it is necessary to switch on between the low-resistance ones by the source of signal and by load (input of amplifier), shunting UA - between the high-impedance ones by the source of signal and the load.

As can be seen from Fig. 62, the semiconductor diode, whose volt-ampere characteristic is described by expression (2-65), is the simplest controlled nonlinear element/cell.

If diagrams of UA are compared with the diagrams of the nonlinear elements/cells, given in Fig. 31, then it is not difficult to see that the shunting attenuators have 1st type volt-ampere characteristics, and consecutive - the 2nd. The form of volt-ampere characteristic of UA can be varied, connecting in series or the in parallel active linear resistors/resistances of different value or applying different combinations of the inclusion/connection of diodes and resistors/resistances.

With the help of the controlled nonlinear elements/cells it is possible to carry out a deeper adjustment in comparison with uncontrolled with the smaller amplitude distortions of signal.

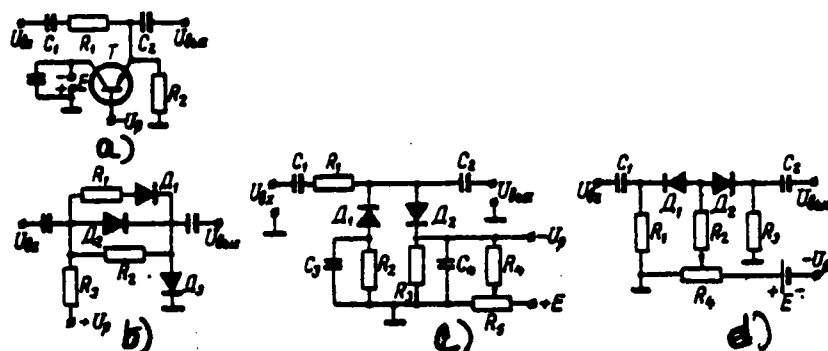


Fig. 62. Diagrams of controlled attenuators: a) asymmetric consecutive; b) asymmetric series-parallel; c) symmetrical shunting; d) symmetrical consecutive.

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For obtaining the smallest distortions, in the first place, the controlled attenuator must be performed with the volt-ampere characteristic, which has a small curvature, in the second place, during the use of consecutive attenuators the adjustment in the amplifier must be realized so that the controlled attenuator would enter voltage/stress the signal of a comparatively small amplitude. If the shunting attenuators are used, the second condition can be excluded.

The transmission factors of shunting $K_{a.m}$ and consecutive $K_{a.c}$

attenuators without taking into account the action of compensating voltages/stresses U_n according to Fig. 63, in which equivalent diagrams of UA are given, are respectively equal to:

$$K_{a. n} = \frac{U_{a n}}{U_{a n}} = \frac{r'_a}{r'_a + R}; \quad (3-80)$$

$$K_{a. n} = \frac{R'}{r'_a + R'}, \quad (3-81)$$

where $\frac{1}{r'_a} = \frac{1}{r_a} + \frac{1}{R_n}$; $\frac{1}{R'} = \frac{1}{R} + \frac{1}{R_n}$; r_a - resistor/resistance of nonlinear element/cell; R_n - resistance/resistor of the load of attenuator.

Let us find the law of a change in control current $I_p = f(z)$, necessary for assigned type realization of FAKh with the help of shunting and consecutive UA. Let us assume that the influencing the nonlinear element/cell signals are sufficiently small. This will make it possible to use the differential parameters of diode for the analysis.

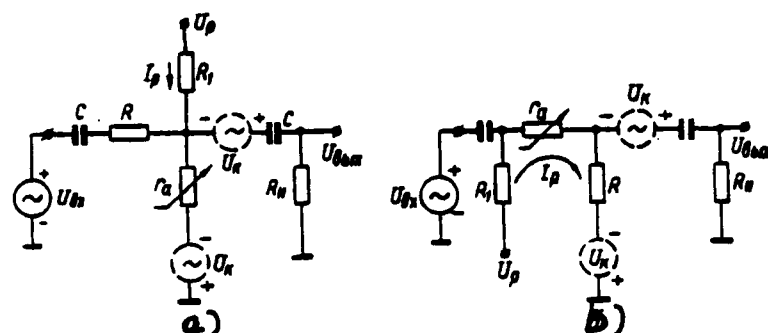


Fig. 63. Equivalent diagrams of the controlled attenuators: a) shunting; b) consecutive.

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Shunting UA.

If as the nonlinear element/cell semiconductor diode with the characteristic, described by expression (2-65) is used, then for the low currents, when inequality $i_{\text{max}} \ll U_{\text{HEH}}$, differential resistor/resistance r_a according to formula (2-66) are fulfilled

$$r_a = \frac{1}{\lambda i_{\text{max}}}. \quad (3-82)$$

Since (Fig. 63) the equality

$$I_p = i_{\text{max}} = A(\exp \lambda U_{\text{HEH}} - 1) = \left(\frac{1}{\lambda r_a} - A \right),$$

is fulfilled then

$$r_a = \frac{1}{\lambda (I_p + A)}. \quad (3-83)$$

Substituting the value r_a into expression (3-80), we obtain

$$K'_a = - \frac{R_H}{R_H + R[1 + \lambda R_H (I_p + A)]}. \quad (3-84)$$

With the sufficiently high currents, when inequality $[\lambda(U_{\text{max}} - i_{\text{max}} r)] \gg 1$, is fulfilled differential resistor/resistance r_d according to (2-68)

$$r_d = r + \frac{1}{\lambda i_{\text{max}}} = r + \frac{1}{\lambda I_p}. \quad (3-85)$$

In this case

$$K_{L, \text{max}} = \frac{R_n (1 + \lambda I_p r)}{R_n (1 + \lambda I_p r) + R [1 + \lambda I_p (r + R_n)]}. \quad (3-86)$$

Taking into account conditions (2-18), (3-84) and (3-86), we obtain expression for control current I_p respectively at the low and high values

$$I'_{p, \text{max}} = \frac{a - \varphi(z) (R_n + R + \lambda R R_n A)}{\varphi(z) \lambda R R_n}; \quad (3-87)$$

$$I''_{p, \text{max}} = \frac{a - \varphi(z) d (R_n + R)}{c \lambda r - \varphi(z) d (R_n r + R r + R R_n)}, \quad (3-88)$$

where

$$\begin{aligned} a &= R [1 + \lambda R_n (I_{p, \text{max}} + A)]; \\ c &= R_n (1 + \lambda I_{p, \text{max}} r) + R [1 + \lambda I_{p, \text{max}} (r + R_n)]; \\ d &= 1 + \lambda I_{p, \text{max}} r; \end{aligned}$$

$I_{p, \text{max}}$ - initial current of adjustment.

Let us determine the maximum depth of adjustment h , which can be carried out with the help of shunting UA. It is important to know parameter h at the development of amplifier-limiters. Maximum transmission factor UA

$$\frac{K_{\text{max}}}{I_p \rightarrow 0} = \frac{R_H}{R_H + R(1 + \lambda R_H A)} \approx \frac{R_H}{R_H + R}, \quad (3-89)$$

since for the real diagrams inequality $\lambda R_H A \ll 1$ is fulfilled.

Minimum transmission factor UA

$$K_{\text{min}} = \frac{r}{r + R + \frac{Rr}{R_H}}$$

Then the maximum depth of the adjustment

$$h_{\text{max}} = \frac{K_{\text{max}}}{K_{\text{min}}} = \frac{R_H(r + R) + Rr}{[R_H + R(1 + \lambda R_H A)]r}$$

If condition $R_H \rightarrow \infty (R_H > R)$, is satisfied then

$$h_{\text{max}} \approx \frac{R+r}{r} \approx \frac{R}{r}, \quad (3-90)$$

where r - volumetric strength of materials of semiconductor, the component of the unit of ohm.

Thus, for obtaining the high value of the coefficient of adjustment it is necessary to in every possible way increase resistor/resistance of R . However, with increase in R maximum transmission factor of divider is reduced. And when $R = R_H$ according

to (3-89)

$$K_{\text{III. MANG}} \approx \frac{1}{2}.$$

For obtaining $K_{\text{III. MANG}} \rightarrow 1$ it is necessary that the condition $R_{\text{II}} \gg R$, $R_{\text{II}} \gg r_{\text{L. MANG}} = \frac{1}{\lambda A}$ would be satisfied

These conditions can be made, if at the output of attenuator is connected the electron-tube cascade/stage, which works without the grid currents (cathode follower is better), or the emitter follower. At the sufficiently high frequencies for this type of transistor even upon the inclusion/connection of the emitter follower the conditions indicated cannot be made and R it is necessary to take finite quantity, on the order of 10-20 kilohm. With $r=10$ ohm we obtain

$$h_{\text{MANG}} \approx (1-2) \cdot 10^3.$$

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For obtaining the high value h_{MANG} it is necessary to compensate for the residual voltage of signal U_{OCT} which is isolated on fixed resistor of r . The value of compensating voltage/stress U_* must be equal to U_{OCT} i.e.

$$\begin{aligned} U_* &= U_{\text{OCT}} = U_{\text{BX}} K_{\text{III. MNG}} = \\ &= U_{\text{BX}} \frac{r}{r+R} \approx U_{\text{BX}} \frac{r}{R}. \end{aligned}$$

The versions of the inclusion/connection of the source of compensating voltage are shown on Fig. 63. During the supplying of

compensating voltage/stress U_n , the voltage of adjustment I_p is equal to zero and value $h_{MENG} \rightarrow \infty$.

The schematic diagram of low-frequency aperiodic amplifier-limiter with the controlled attenuator on the output and the compensating voltage/stress, which is removed/taken from resistor/resistance of R_s , connected with the emitter circuit of amplifier, and is supplied to the nonlinear element/cell through phase-shift circuit R_s, C_s, R_s, C_s , it is depicted in Fig. 64. At frequencies $f=50 \text{ Hz} \dots 20 \text{ kHz}$ the coefficient of adjustment is obtained to $(5-8) \cdot 10^3$.

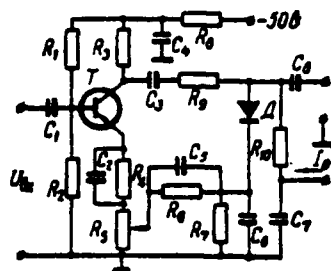


Fig. 64. The schematic diagram of amplifier-limiter with the controlled shunting attenuator at the input: T_1 - P423A; D_1 - D233B; C_1, C_3, C_5 - $5 \mu\text{g}$; C_2, C_4 - $29 \mu\text{F}$; C_6 - $0.01 \mu\text{F}$; C_7 - $0.06 \mu\text{F}$; R_1 - 15 kilohm; R_2 - 6.8 kilohm, R_3 - 1.5 kilohm; R_4 - 2.7 kilohm; R_5 - 100 ohms; R_6 - 18 ohms; R_7 - 33 ohms; R_8 - 3.3 kilohm; R_9 - 4.7 kilohm; R_{10} - 10 kilohm.

Consecutive UA (Fig. 63b).

Upon the series connection of nonlinear element/cell the transmission factor consecutive UA is maximum with the high currents of adjustment. According to expressions (3-84) and (3-85) the transmission factor of attenuator with the high currents

$$K_{a,a}^* = \frac{\lambda R' I_p}{1 + \lambda R' (I_p + R')} ; \quad (3-91)$$

respectively with low currents I_p

$$K_{a,a}^* = \frac{R' \lambda (I_p + A)}{1 + R' \lambda (I_p + A)} . \quad (3-92)$$

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Taking into account condition (2-18), we obtain the required law of a change in current I_p for the realization of FAKh

$$I_{p. n}'' = \frac{\varphi(z) I_{p. n}}{c - \varphi(z) \lambda I_{p. n} (r + R')}; \quad (3-93)$$

$$I_{p. n}' = \frac{\frac{z}{x} [b(1 + A)] - Aa}{a - bR' \lambda \frac{z}{x}}, \quad (3-94)$$

where

$$a = 1 + R' \lambda (I_{p. n} + A); \quad b = I_{p. n} + A; \quad c = 1 + \lambda I_{p. n} (r + R').$$

Transmission factor is maximum when $I_p \rightarrow \infty$

$$K_{n. max} = \frac{R'}{r + R'} \approx 1, \quad (3-95)$$

since $R' \gg r$.

Transmission factor is minimal when $I_p \rightarrow \infty$

$$K_{n. min} = \frac{R' \lambda A}{1 + R' \lambda A} \approx R' \lambda A, \quad (3-96)$$

since for the real diagrams is fulfilled inequality $R' \lambda A \ll 1$. Maximum depth of the adjustment

$$h_{n. max} = \frac{K_{n. max}}{K_{n. min}} = \frac{1}{R' \lambda A}. \quad (3-97)$$

As can be seen from expression (3-97), for an increase in the depth of adjustment it is necessary to apply diodes with small

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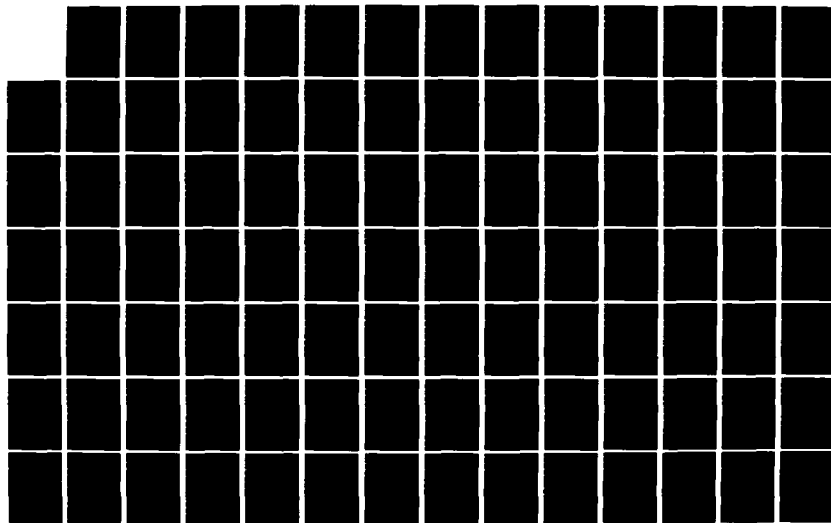
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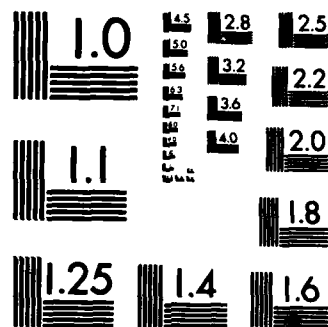
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inverse currents (small parameter A) and to switch on UP to low-resistance load. If load U_A is the input resistance of transistor

$R_{in} \approx 10^6$ ohm, then with $A = (10^{-4} - 10^{-7})$ a

$$h_{u, max} = 2,5(10^4 + 10^6).$$

Thus, with the low-resistance load with the help of consecutive U_A it is possible to realize the very deep adjustment of transmission factor. On the radio frequencies with an increase in the frequency the depth of adjustment considerably descends due to the effect of the transfer capacitance of nonlinear resistance and stray capacitances.

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APPLICATION OF INTERSTAGE COUPLING ELEMENTS WITH THE OPTIMUM PARAMETERS.

The instability of frequency and phase characteristics of aperiodic and selective amplifiers over the dynamic range can be considerably reduced, applying equivalent components of the interstage connection/communication, which let us agree to call optimum.

To such elements/cells can be attributed nonlinear dividers for the aperiodic amplifiers and single oscillatory circuit and

two-circuit band-pass filter with the optimum parameters for the selective amplifiers.

From the enumerated elements/cells not examined is only the single oscillatory circuit. Present paragraph is dedicated to the analysis of single oscillatory circuit with the optimum parameters.

The equivalent diagram of parallel circuit is given in Fig. 65.

The general/common/total conductivity of duct/contour is determined by the relationship/ratio

$$Y_0 = \frac{\omega^2 L^2 g_1 R_1^2 + R_1}{R_1^2 + \omega^2 L^2} + j \frac{R_1^2 \omega C + \omega^3 L^2 C - \omega L}{R_1^2 + \omega^2 L^2}, \quad (3-98)$$

where $g_1 = 1/R_1$.

The condition of parallel resonance it is

$$I_m Y_0 = 0, \quad (3-99)$$

or in accordance with expression (3-98)

$$R_1^2 C + \omega^2 L^2 C - L = 0.$$



Fig. 65. The equivalent diagram of the parallel oscillatory circuit: C and L - capacity/capacitance and the inductance of duct/contour; R_2 - resistor/resistance of insertion losses and losses in the capacity/capacitance; R_1 - resistor/resistance of losses in the inductance.

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Whence we find the resonance frequency of the system

$$\omega_1 = \sqrt{\frac{L - CR_1^2}{LC}} = \omega_0 \sqrt{1 - \frac{d^2}{1}} = \omega_0 \sqrt{1 - d^2}, \quad (3-100)$$

where $\delta = R/2L$ - decrement of circuit damping; $d = 1/Q = R_1/\rho$ - circuit damping; $\omega_0 = \frac{1}{\sqrt{LC}}$.

The conductance of the duct/contour

$$g = \frac{g_0 \omega^2 L^2 + R_1^2 g_1 + R_1}{R_1 + \omega^2 L^2}.$$

Substituting the value $\omega = \omega_1$, we obtain expression for the conductance in the presence of the resonance

$$g_0 = \frac{g_0 \omega_0^2 (1 - d^2) L^2 + g_1 + g_1}{1 + \frac{\omega_0^2 L^2}{R_1^2} (1 - d^2)} = g_1 + \frac{d}{\rho}.$$

Using equality (3-98), we find expression for the phase response

$$\varphi = \operatorname{arctg} \frac{b}{g} = \operatorname{arctg} \frac{m\eta(d^2 + \eta^2 - 1)}{\eta^2 + d^2 + dm}, \quad (3-101)$$

where $\eta = \omega/\omega_0$; $m = R^2/\rho$.

Relationship/ratio (3-101) is precise and determines phase displacement at any frequency. It is obvious that at the resonance frequency phase displacement is equal to zero, and on the band edges of transmission - respectively $\pm\pi/4$.

Actually/really, resonance frequency is determined by relationship/ratio (1-100). Then

$$\eta_1 = \frac{\omega_1}{\omega_0} = \sqrt{1-d^2}; \quad d^2 + \eta^2 - 1 = 0 \Rightarrow \varphi = 0.$$

We know that circuit susceptance at the edge of the passband $b = \pm g$. Then

$$\varphi = \operatorname{arctg} \left(\pm \frac{b}{g} \right) = \operatorname{arctg} (\pm 1) = \pm \frac{\pi}{4}.$$

Let us find the divergences of phase characteristic with a change in the parameters of duct/contour. In this case we will consider the fact, usually not specified and to literature and on which, as a rule, they do not accentuate attention. With a change in the argument to value Δx function $f(x)$ at the assigned point x , obtains the increment

$$\Delta f(x_0, \Delta x) = f(x_0 + \Delta x) - f(x_0). \quad (3-102)$$

Usually value Δf is determined through total differential, substituting infinitesimal increments df and dx by finite increments $\Delta f(x)$ and Δx , i.e., they count

$$\Delta f_1(x_0, \Delta x) = f'(x_0) \Delta x. \quad (3-103)$$

Value $\Delta f_1(x_0, \Delta x)$ at any value Δx is equal to true increment $\Delta f(x_0, \Delta x)$, if function $f(x)$ is linear. Actually/really, $f(x) = ax + b$ and $f(x_0) = ax_0 + b$. Then

$$\begin{aligned} \Delta f(x_0, \Delta x) &= f(x_0 + \Delta x) - f(x_0) = [a(x_0 + \Delta x) + b] - \\ &\quad - (ax_0 + b) = a\Delta x. \end{aligned}$$

On the other hand, $f'(x_0) = f'(x) = a$ and, therefore,

$$\Delta f_1(x_0, \Delta x) = f'(x_0) \Delta x = a\Delta x.$$

But if $f(x)$ nonlinear function, then the values, determined from expressions (3-102) and (3-103), in the general case are not equal. The difference between them vanishes with $\Delta x \rightarrow 0$ and in the general case it is determined by the convexity (concavity) of function. If curved convex, which corresponds to $f''(x) > 0$, then the result, determined from relationship/ratio (3-103), proves to be understated, while if concavity is converted down, which corresponds to $f''(x) < 0$, then result is overstated. Difference in the results, i.e., the error in relationship/ratio (3-103) is greater, the greater the value $f''(x_0)$. Moreover with the actually permissible changes in the parameters 10-20% error sometimes can reach the significant magnitude. Therefore let us consider error during the determination of the value of a change in the phase response on the band edge of

transmission.

Let $f(x) = \text{arctg } x$; $x_0 = \pi/4 = 0.8$. Then

$$\Delta f(x_0, \Delta x) = \text{arctg}(x_0 \pm \Delta x) - \text{arctg } x_0 = \text{arctg} \frac{\Delta x}{1 + x_0^2 \pm \Delta x x_0};$$

$$\Delta f_1(x_0, \Delta x) = f'(x_0) \Delta x = \frac{\Delta x}{1 + x_0^2};$$

Let $\Delta x = 0.1$. Then

$$\Delta f(x_0, \Delta x) = 0.058; \quad \Delta f_1(x_0, \Delta x) = 0.061.$$

Hence relative error Δf_1 in comparison with Δf composes 6%.

Let $\Delta x = 0.4$. Then $\Delta f(x_0, \Delta x) = 0.3$; $\Delta f_1(x_0, \Delta x) = 0.24\%$ and a relative error in the calculation it composes 20%.

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Thus, with considerable relative changes in the argument an error in the determination of divergence ϕ with respect to approximation formula (3-103) can reach the significant magnitude.

In the transistor amplifiers, which have considerable changes in their own parameters, the changes of the phase ϕ , calculated by formula (3-103), can be overstated. Therefore for the functional amplifiers, especially transistor, a change in the phase should be determined from precise formula (3-102).

Let us find the dependences of change φ on a change in the parameters of duct/contour. In the oscillatory system in question by variable quantities can be C_1 and R_1 . Let $C_1 = C_0 + \Delta C$ and $R_1 = R_0 + \Delta R_1$. Then

$$\begin{aligned}\omega_0 + \Delta\omega &= \frac{\omega_0}{\sqrt{1 + \frac{\Delta C}{C_0}}}; \\ \eta_0 + \Delta\eta &= \eta_0 \sqrt{1 + \frac{\Delta C}{C_0}}; \\ d_0 + \Delta d &= d_0 \sqrt{1 + \frac{\Delta C}{C_0}}; \quad m_0 + \Delta m = m_0 \left(1 + \frac{\Delta R_1}{R_1}\right) \sqrt{1 + \frac{\Delta C}{C_0}},\end{aligned}$$

since $m = R_1 / \rho$ and $\rho_0 + \Delta\rho = \frac{\rho_0}{\sqrt{1 + \frac{\Delta C}{C_0}}}$.

Hence precise formula for determining the change in the phase response of system with a change in the parameters

$$\begin{aligned}\Delta\varphi &= \text{arctg} \frac{\eta m_0 \left(1 + \frac{\Delta R_1}{R_1}\right) \left[d_0^2 \left(1 + \frac{\Delta C}{C_0}\right) + \eta^2 \left(1 + \frac{\Delta C}{C_0}\right) - 1 \right]}{\eta^2 + d_0^2 + d_0 m_0 \left(1 + \frac{\Delta R_1}{R_1}\right)} - \\ &\quad - \text{arctg} \frac{\eta m_0 (d_0^2 + \eta^2 - 1)}{\eta^2 + d_0^2 + d_0 m_0}.\end{aligned}\quad (3-104)$$

Approximate relationship/ratio, obtained analogously with expression (3-103) and making it possible to determine change φ with $0 \leq |\Delta C / C_0| \leq 0.2$,

$$\Delta\varphi_G = \frac{\Delta z_1}{1 + z_1^2}, \quad (3-105)$$

where

$$z = \frac{\eta m_0 (d_0^2 + \eta^2 - 1)}{\eta^2 + d_0^2 + d_0 m_0};$$

$$\Delta x_1 = \frac{(\Delta m \eta + m_0 \Delta \eta) (d_0^2 + \eta^2 - 1) + m_0 \eta (2d \Delta d + 2\eta \Delta \eta)}{\eta^2 + d_0^2 + d_0 m_0} -$$

$$- \frac{m_0 \eta (d_0^2 + \eta^2 - 1) (2\eta \Delta \eta + 2d \Delta d + m_0 \Delta d + \Delta m d_0)}{(\eta^2 + d_0^2 + d_0 m_0)^2};$$

$$\Delta \eta = -\frac{1}{2} \eta_0 \frac{\Delta C}{C_0}; \quad \Delta d = -\frac{1}{2} d_0 \frac{\Delta C}{C_0};$$

$$\Delta m = \frac{1}{2} m_0 \frac{\Delta C}{C_0}.$$

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After simple, but bulky conversions, we obtain

$$\Delta \varphi_C = \frac{\Delta C}{C_0} \cdot \frac{m_0 \eta (d_0^2 + \eta^2) (d_0 m_0 + 1)}{(\eta^2 + d_0^2 + d_0 m_0)^2 + m_0^2 \eta^2 (d_0^2 + \eta^2 - 1)^2}. \quad (3-106)$$

Is analogous with change R, change φ with $0 < |\Delta R_1 / R_{10}| \leq 0.2$

$$\Delta \varphi_{R_1} = \frac{\Delta x_1}{1 + x^2}, \quad (3-107)$$

where

$$\Delta x_1 = \frac{\eta (d_0^2 + \eta^2 - 1) \frac{\Delta R_1}{R_1}}{\eta^2 + d_0^2 + d_0 m_0} - \frac{d_0 \eta m_0 (d_0 + \eta^2 - 1) \frac{\Delta R_1}{R_1}}{(\eta^2 + d_0^2 + d_0 m_0)^2}.$$

After conversions

$$\Delta \varphi_{R_1} = \frac{\Delta R_1}{R_1} \cdot \frac{\eta (d_0^2 + \eta^2 - 1) (\eta^2 + d_0^2)}{(\eta^2 + d_0^2 + d_0 m_0)^2 + m_0^2 \eta^2 (d_0^2 + \eta^2 - 1)^2}. \quad (3-108)$$

With the simultaneous change C and R, the value of change φ is equal to the sum of the changes, determined to relationships/ratios (3-106) and (3-108), i.e.

$$\Delta \varphi = \Delta \varphi_C + \Delta \varphi_{R_1}. \quad (3-109)$$

Let us find the condition of the stabilization of the phase response of duct/contour with change in the capacity/capacitance. It is obvious that the phase response is not changed, if

$$\frac{\partial \varphi}{\partial C} = 0. \quad (3-110)$$

Taking into account that in this case the functions of capacity/capacitance C are η , m and d , according to expressions (3-101) and (3-110)

$$[(m'\eta + m\eta')(d^2 + \eta^2 - 1) + m\eta(2dd' + 2\eta\eta')](\eta^2 + d^2 + dm) - (2\eta\eta' + 2dd' + dm' + d'm)m\eta(d^2 + \eta^2 - 1) = 0. \quad (3-111)$$

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Since

$$\eta' = \frac{\partial \eta}{\partial C} = -\frac{1}{2} \eta_0 \frac{\partial C}{C}; \quad d' = \frac{\partial d}{\partial C} = -\frac{1}{2} d_0 \frac{\partial C}{C}; \quad m' = \frac{\partial m}{\partial C} = \frac{1}{2} m_0 \frac{\partial C}{C},$$

after conversions we obtain

$$md = 1. \quad (3-112)$$

Substituting the value of m and d , expressed through ρ , we obtain

$$\frac{R_2}{\rho} = \frac{\rho}{R_1},$$

or

$$R_1 R_2 = \rho^2. \quad (3-113)$$

If one considers that $\rho=1/\omega.C.$, then the condition for a minimum change in the phase ϕ with a change in capacity/capacitance C can be registered in the following form:

$$R_2\omega C = \frac{1}{R_1\omega C}. \quad (3-114)$$

Relationships/ratios (3-113) and (3-114) coincide with the conditions, obtained by S. A. Shkabara, on the basis of energy relationships/ratios [72].

Condition (3-114) structurally/constructurally can be performed by two methods: connecting in series with the inductance of duct/contour resistor/resistance R_1 , manufacturing self-inductor from the wire with the high ohmic resistance.

The curves of a change in the energy factor of the usual and stabilized duct/contour are given in Fig. 66. From the curves it is evident that the effect of stabilization is considerable with a change in the capacity/capacitance of duct/contour to 50%.

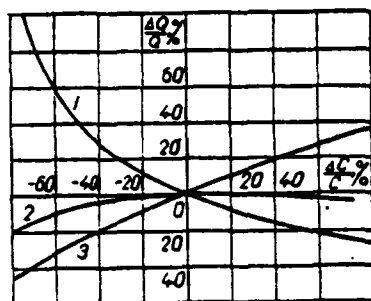


Fig. 66. The curves of a change in the energy factor of oscillatory circuit with a change in the capacity/capacitance: 1 - $R_1 > \rho^2/R_2$; 2 - $R_1 = \rho^2/R_2$; 3 - $R_1 < \rho^2/R_2$.

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Chapter 4.

FUNCTIONAL AMPLIFIERS WITH THE ADDITION OF OUTPUT EFFECTS.

51. Conditions of the construction of amplifier circuits.

Output effects independent of circuit solution and structure of the reinforced signal must store/add up on the total load linearly, otherwise the AKh of the amplifier will differ from functional. Linear addition can be realized, if the source of output effect is current generator with a high internal resistor/resistance of R_r (Fig. 67), signals in the form of video pulses synchronize and polarity, and signals in the form of radio oscillations coincide in the phase.

For satisfaction of the first condition must be performed the inequality

$$R_r > 10R_{\Sigma o.} \quad (4-1)$$

where $R_{\Sigma o.}$ — the total resistance of the load of current generators.

Current generator in the diagram in Fig. 67 substitutes amplifier stage for aperiodic FU or nonlinear amplifier nucleus, which consists of amplifier stage and corrective element/cell (detector, the untying cascade/stage) for the selective amplifiers, and is considered the delay time of the pulse signal t , or the change in the phase of radio signal $\Delta\phi$. For the coincidence of the fronts of video pulses and obtaining of the identical phase of the radio signals, which enter the total load, it is necessary to apply the special time-delay or phase-correcting devices/equipment.

For consecutive type aperiodic and sufficiently narrow-band selective pulse FU with the video output for the coincidence of the fronts of video pulses it is possible to apply artificial delay lines, which must satisfy the following requirements:

1. The delay time of the segment of line, connected between two amplifier stages, in the first approximation, must equal the transit time of the impulse/momentum/pulse through amplifier stage $t_{a.m}$ for the aperiodic amplifier

$$t'_{a.m} = C_0 R_0; \quad (4-2)$$

for the single-circuit selective amplifier

$$t''_{a.m} = 2C_0 R_0. \quad (4-3)$$

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2. Delay lines must have sufficiently wide passband so that video pulses would not be distorted. The wider the passband of line, the lower the rise time and distortion of video pulse at the output of line.

3. Delay lines must have high quality in order not to introduce high attenuations.

During timing of delay according to formula (4-2) it is necessary to remember that in the control of amplification the passband of cascade/stage is changed, as a rule, is widened. When the very high accuracy of the coincidence of the pulse edges, which are folded on the total load, is required, in formula (4-2) it is necessary to substitute the value the passband of cascade/stage, which corresponds to the average/mean value of signal level at the input of this cascade/stage. The best results are obtained, if the value of passband is taken with the signal, which corresponds to the middle of the functional section of AKh of cascade/stage.

Since FU is nonlinear, during the amplification of pulse signals during stricter calculations should be determined delay time on the base of the transient processes, which take place in amplifier stage,

on different input signal levels.

A detailed theory of artificial delay lines (filters) with concentrated and distributed parameters, and also the procedure of calculation of these lines are presented in works [46, 86]. In the present work only some positions about the delay lines, necessary for designing the amplifiers with FAKh, are examined. Delay lines with concentrated constants can consist of T- and U-shaped components/links of the type of the constant k or of the derived components/links of the type m .

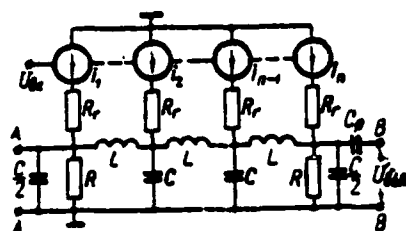


Fig. 67. Equivalent diagram of FU with the addition of output effects.

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In the equivalent diagram, depicted in Fig. 67, the delay line with concentrated parameters, which consists of the U-shaped components/links of the type k, is used. The delay time of a component/link of the type k is determined from the formula

$$t_k = \sqrt{LC}, \quad (4-4)$$

where L - series inductance of component/link; C - shunt capacitance of component/link.

If delay line consists of n components/links, then the caused on time base of the delay

$$T_s = nt_s, \quad (4-5)$$

If between the amplifiers of cascade/stage it is connected on one component/link, then must be performed the equality

$$t_s = t_k. \quad (4-6)$$

Any section of delay line, which in the case of selective amplifiers combines the detectors of adjacent cascades/stages, it is the at the same time interstage filter, which does not pass intermediate frequency. So that in the delay line reflections would not appear, it must be loaded from the output and the input with the effective resistance, equal to its characteristic (wave) resistor/resistance

$$Z \approx Z_{\omega} = \rho = \sqrt{\frac{L}{C}}, \quad (4-7)$$

where ω - angular frequency.

One should note that the delay time t , and line characteristic depend on frequency. Therefore virtually it is not possible to match line with the load in the wide passband of frequencies.

In work [44] it is shown that only for one section of the type k are found the optimum values of load for the case, when load consists of effective resistance of R , which do not depend on the frequency. With attains the constancy only of one of the parameters of delay line in the broadband. For example, in order to obtain active output resistance of line for a T- component/link, one should choose $R=0.75\rho$ in order to obtain the minimum of reactive component of the input resistance $R=0.95\rho$, and the constancy of delay time - $R=0.97\rho$. For a

P- component/link the values of these coefficients are respectively equal to: 1.5; 2.06; 1.65.

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As a result of a change in the characteristic impedance to match delay line with the load is virtually possible only in the part of the band of transparency. For the components/links of the type k the characteristic impedance is approximately constant in the limits approximately/exemplarily of half of the band of transparency. Therefore during calculations of delay lines, which consist of the components/links of the type of the constant k, even when stringent requirements on the accuracy of reproduction of impulses/momenta/pulses are not imposed, tentatively it is possible to count the passband of frequencies not more than half of the band of transparency. The poor use of a band of transparency in the delay lines, comprised of the components/links of the type of the constant k, is essential deficiency/lack and limits their use/application.

The band of transparency considerably better is used, if derived components/links of the type m are applied, in this case it is possible to widen the frequency band, in limits of which is realized the agreement with the load, and the frequency band, within limits of which sufficiently accurately is retained the value of delay time; to

increase the delay time, caused on separate component/link, without a change in the pulse rise-time at the output.

Delay time of a derived component/link of the type m [46]

$$t_s' = m \sqrt{LC} \quad (4-8)$$

Value m, at which the delay time remains constant in possible broadband, is equal to 1.23 [44]. To even more decrease the frequency and phase distortions in the delay lines is possible, connecting in parallel to the inductive elements of delay line special corrective capacities/capacitances [46].

Very good data have delay lines with the distributed parameters, made with the broad band transparency (with the maximum cut-off frequency to 10-20 MHz) and with the high quality. It is most expedient to apply such lines in UPCh, intended for amplifying the narrow pulses with a duration of $t_n < 0.5 \mu s$.

However, with an increase in the maximum boundary-like character the frequencies of the delay line of the condition of the decoupling between the cascades/stages deteriorate and amplifier is inclined to the self-excitation.

For the broadband selective amplifiers, when the inequality

$$\frac{\pi}{f_0} > 0,5 \quad (4-9)$$

is fulfilled and the delay time of cascade/stage very little, to apply delay line as the untying element/cell there is no sense. In this case should be applied the untying cascades/stages, which can be connected in parallel and loaded to the general/common/total effective resistance (Fig. 16e) or connected parallel-series (Fig. 16g). In this case the pulse edges on the time do not coincide. If inequality $t_{\text{d}} \ll t_{\text{p}}$ is fulfilled the addition of impulses/momenta/pulses is realized sufficiently well.

If at both ends/leads of the delay line are connected the effective resistance, equal in magnitude to the line characteristic of delay $\rho = R$ (Fig. 67), the total resistance of load on the video frequency for all current generators $R_{\text{L.0}} = 0,5\rho$. Then according to (4-1) at the given value of the internal resistor/resistance of current generator R_r resistor/resistance ρ must be chosen from the condition

$$\rho < \frac{R_r}{5}. \quad (4-10)$$

S2. Aperiodic functional video amplifiers.

In the aperiodic amplifiers of UP in essence they switch on by circuits with the common cathode or the general/common/total emitter,

which change the phase of the reinforced signal on 180° . Because of this it is not possible to sum the pulse signals, taken directly with the output of UK.

For adding the video pulses in consecutive type multistage FU it is necessary at the outputs of odd (or even) amplifier stages to include/connect the phase-inverting cascades/stages (FIK), which turn the phase of signal on 180° . If FIK are connected at the outputs of odd cascades/stages (elements/cells 1', 3', ... in Fig. 16c), then for the decoupling of remaining cascades/stages and guarantee of stable work of amplifier at the outputs of even UK it is necessary to include/connect the cathode (KP) or emitter (EP) followers (elements/cells 2', 4' ... in Fig. 16c).

As FIK can be used the cascade/stage, made on the diagram with the common cathode or the general/common/total emitter. Amplitude characteristic of FIK must be linear in ShDD. For this in FIK is introduced deep negative feedback.

The schematic diagram of two cascades/stages aperiodic pulse of FU on the tubes is shown in Fig. 68.

At the output of the first cascade/stage, assembled on the tube L_1 , is connected FIK, made on the tube L_2 ; at the output of the second cascade/stage (L_2) it is connected the cathode follower, assembled on the tube L_3 . As can be seen from diagram in Fig. 68, the resistor/resistance of current generator on the circuit of FIK is R_1 , and on the circuit KP - sum of resistors/resistances $R_1 + R_2$. According to the condition

$$(4-1) R_1 > R_{n.o} = \frac{\rho}{2}.$$

Then transmission factor on the circuit FIK

$$\begin{aligned} k'_p &= \frac{S R_{n.o}}{1 + S_n(R'_1 + R_2)} = \\ &= \frac{S \rho}{2[1 + S_n(R'_1 + R_2)]}, \quad (4-11) \end{aligned}$$

where S_n - the slope/transconductance of the cathode characteristic of tube; R'_1 - resistor/resistance of feedback in the cathode circuit of tube L_3 .

Transmission factor on the circuit KP when $R_1 + R'_1 \gg R_{n.o}$

$$k''_p \approx \frac{S_n R_{n.o}}{1 + S_n(R_1 + R'_1)} = \frac{S_n \rho}{2[1 + S_n(R_1 + R'_1)]}. \quad (4-12)$$

Transmission factors k_p and k''_p can be changed, changing the depth of negatively current feedback of signal with the help of the slides/wipers, blocked by great capacities of C_1 and C_2 . With a change in the values of the resistors/resistances of feedback R'_1 and

R' , in the signal current the values of resistors/resistances for the direct current in the cathodes of tubes L_1 and L_2 remain constant/invariable. This causes the constancy of the mode of operation of tubes L_1 and L_2 on the direct current during the adjustment of transmission factors k'_p and k''_p .

For obtaining the assigned form of FAKh the coefficients of transmission k'_p and k''_p in the multistage amplifier must be calculated by the formulas, presented in §5 of Chapter 1.

Advantages of diagram.

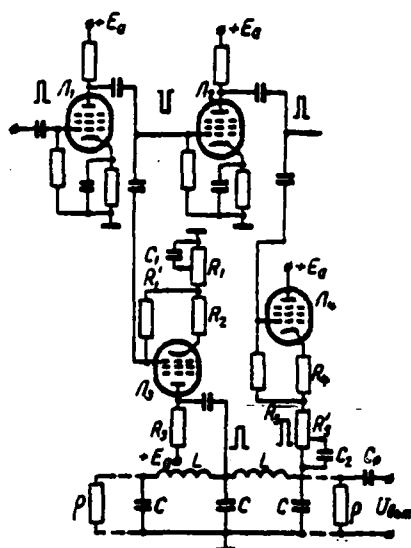


Fig. 68. Schematic diagram of logarithmic video amplifier with the addition of output effects.

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Possibility of realization of FAKh with $b = \text{var}(\downarrow)$ and (\uparrow) , i.e., any form.

Possibility of obtaining the sufficiently stable delay time total output pulse with the execution of amplifier with the diagram (Fig. 16g) without the delay line. This is possible to achieve, selecting the appropriate value of the time constant of loads of FIK (elements/cells 1", 3", ... on Fig. 16g) and K_P (elements/cells 2",

4", ... in Fig. 16g).

However, diagram has the following deficiencies/lacks:

Diagram is fairly complicated.

On the high signal level appears the considerable parasitic reverse/inverse overshoot, which toward the end of the dynamic range of FAKh with $b_{\text{var}}(\downarrow)$ of 60-70 dB can reach 50-80% of the value of the signal pulse in the dependence on the selected modes/conditions of the tubes of amplifier stages. After the break-down of the signal pulse the diagram for a prolonged time loses sensitivity.

This phenomenon is caused by the appearance of considerable grid currents in the cascades/stages, on input of which the impulses/momenta/pulses of positive polarity act.

Everything said relative to diagram on the tubes in the equal measure relates also to the transistor circuit.

From the deficiencies/lacks indicated are to a considerable extent free FU, in which UK are made on the cascode diagrams: the general/common/total anode - common grid (OA-OS) or common collector/receptacle - general/common/total base (OK-OB). Such

cascades/stages have high input resistance and is not changed the phase of the reinforced signal.

The schematic diagram of the logarithmic amplifier, made on the cascode diagrams OA-OS, is given in Fig. 69. Let us consider the first cascade/stage, assembled on the tubes L_1 and L_2 .

For obtaining FAKh of multistage amplifier it is possible to summarize the signals, taken with any of three points 1, 2 or 3. However, it is most expedient to summarize the signals, taken from the anode of the first tube L_1 (points 1),. In this case the virtually complete decoupling of the circuit of addition and circuit of amplification without the further untying cascades/stages is obtained. During the addition of the signals, taken from points 2 or 3, it is necessary to place the further untying cascades/stages, which is economically disadvantageous.

From the diagram in Fig. 69 it is evident that the first part of the cascode diagram, made on the tube L_1 , is phase-inverting cascade/stage of FIK, and the second part on the tube L_2 - by amplifier.

For simplification in the diagram the anodes of the tubes of phase-inverting cascades/stages (L_1, L_2, \dots, L_{11}) are connected directly to the resistance/resistor of load R_1 , without the delay line. It is natural that in the diagram in question also can be used the delay line.

To the input of diagram it is necessary to supply the pulse signal of negative polarity. Then on the total load the impulses/momenta/pulses of positive polarity will store/add up. With the increase of signal into the mode/conditions of limitation by the first enters latter/last FIK (L_{11}), then next-to-last, etc. Limitation in the phase-inverting cascades/stages is realized due to the cutoff of anode current in the absence of grid currents. Let us assume that with the closing of tube of FIK circuital current of the anode will be changed to value $\Delta I_a = I_{a.k}$. Then the maximum output voltage of signal taken from total load $R_{n.o}$.

$$U_{\text{max. maxc}} = R_{n.o} \sum_{i=1}^n \Delta I_{ai}, \quad (4-13)$$

where n - quantity of FIK that entered the cutoff conditions.

For the realization of the linear addition of voltages/stresses on the total load must be performed the inequality

$$U_{a. \text{min}} = E_a - U_{\text{max. maxc}} = E_a - R_{n.o} \sum_{i=1}^n \Delta I_{ai}, \quad (4-14)$$

where $U_{a. \text{min}}$ - minimum voltage on the anode of tube of FIK, at whom the parameters of tube still virtually are not changed.

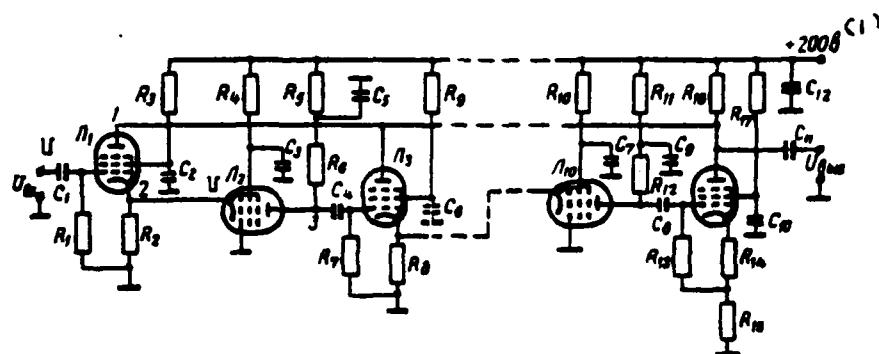


Fig. 69. Schematic diagram of the logarithmic video amplifier with the addition of voltages/stresses on the cascode diagrams.

Key: (1). +200 V.

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Cascode diagram can be assembled on the twin triode, included in one tank/balloon, which reduces the overall sizes of amplifier [43]. However, the parameters of triode to a considerable extent are changed with a change in the voltage on the anode of tube, in consequence of which on the high signal levels the linearity of the addition of signals is broken. On the reason indicated cascode diagrams must be performed on the pentodes.

With identical tubes and satisfaction of condition $R_1 > 1/S$ it is possible to consider that FIK on the cathode circuit is loaded to the

input resistance of cascade/stage with the common grid, which is approximately/exemplarily equal to $R_{in} = \frac{1}{S}$. Then the coefficient of transmission of FIK on the cathode circuit with the identical tubes tentatively can be accepted

$$K_{\phi. n} = \frac{S \cdot \frac{1}{S}}{1 + S \frac{1}{S}} \approx 0,5.$$

Choosing the appropriate initial mode of operation of tubes on the direct current, it is possible to carry out their work without the grid currents in entire dynamic range of FAKh. It is necessary to note that the appearance of grid currents in the second tube of cascode diagram is not dangerous, since between the tubes there is a conductive coupling and isolating capacitor is absent.

On the diagram, analogous to Fig. 69, logarithmic transistorized amplifier can be made. If amplifier is made on transistors of type p-n-p, on the input of amplifier it is necessary to supply the impulses/momenta/pulses of positive polarity, and vice versa, if transistors of type n-p-n are used, to the input of amplifier should be supplied the impulses/momenta/pulses of negative polarity.

In the amplifiers, made on the cascode diagrams, it is possible to obtain LAKh of sufficiently high accuracy. This is explained by the fact that general/common/total AKh of amplifier is formed/shaped due to the amplitude characteristics of phase-inverting

cascades/stages. The latter, in turn, are formed/shaped with the lower section of passage characteristic of UP, which sufficiently accurately (especially in the transistors) reproduces logarithmic law with the closing of UP.

The five-stage amplifier, whose schematic is given in Fig. 69, has the following data: $K_v=9400$; upper cut-off frequency with the work in linear conditions $F_{max}=1.3$ MHz; $D_{max}=70$ dB; $U_{s.s.} = 4 \cdot 10^{-4}$ V; $\delta U=4\%$ converted slope/transconductance of LAKh $\sigma=0.4$ V/dB.

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Slope/transconductance of LAKh it is possible to vary in the wide repartitions/conversions, changing the load on FIK.

In the amplifiers, made on the transistors, to one cascade/stage the range of LAKh to 20 dB can be obtained. Slope/transconductance of LAKh is less than in the vacuum-tube amplifiers.

The absence of parasitic reverse/inverse overshoots and the virtually instantaneous restoration/reduction of the sensitivity of diagram after the break-down of large pulse signal is one of the essential advantages of the logarithmic video amplifier, made on the cascode diagrams.

The need of applying the increased feed on the circuits of the collector/receptacle of phase-inverting cascades/stages is certain structural/design deficiency/lack, which is inherent in all transistor circuits with the addition of output voltages/stresses.

\$3. Consecutive type functional selective amplifiers with the video output.

In the selective amplifiers both on the tubes and on the transistors it is easiest to obtain FAKh on the video voltage, after forming the detected voltages/stresses on the total load. Sufficiently many works are devoted to questions of the circuit solution of the logarithmic amplifiers with the consecutive addition of the detected voltages/stresses. The essence of the work of all diagrams is identical (\$5, Chapter 1). Circuit solutions are characterized by a quantity of instruments, utilized for amplification and detection of radio signals. To reinforce and to detect the voltage/stress of high frequency can one or two separate electronic devices. Depending on this of the schematic of functional selective amplifiers with the video output it is possible to divide into the diagrams: with the separate detectors; with the cathode or emitter detection; by anodic or collector detection; by grid or base

detection.

Functional amplifiers with the separate detectors.

The simplified diagram of FU with the separate detectors is given in Fig. 70. In the given diagram the voltage/stress of high frequency is reinforced by cascades/stages V_1, \dots, V_{n-1}, V_n to outputs of which they are connected detectors A_1, \dots, A_{n-1}, A_n having particular loads R_n and general/common/total $R_{\Sigma} = \frac{1}{2}$, on which they store/add up the detected voltages/stresses.

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Each UP works as amplifier or as the output stage. With the small input voltage the cascades/stages have linear amplification, then with an increase in the voltage/stress amplification is reduced and finally cascade/stage is saturated. Since with the increase of signal all cascades/stages, beginning from the latter, consecutively/serially are overloaded, it is very important that the behavior of cascade/stage in the handled state would be completely specific. It is necessary also that output potential of cascade/stage with saturation would be constant and it did not depend on the value of signal.

During the development of diagram of FU with the separate detectors it is necessary to know, what diagram of UK should be used; it is necessary whether to apply in the diagram of ARU, based on the nonlinearity of characteristic of UP; to what point of diagram of UK most expedient to connect detector.

In selective FU with the separate detectors it is possible to apply all known circuit diagrams of UP, including cascode. Considerations by choice of diagram in essence are the same as for the linear amplifier.

Functional amplifier can be made both from ARU, based on the nonlinearity characteristic of UP and without it.

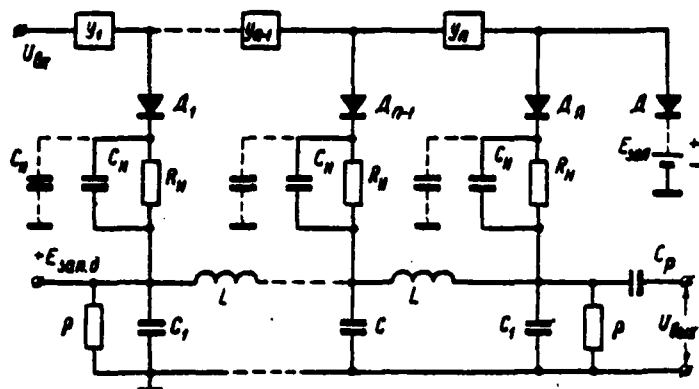


Fig. 70. Simplified circuit of FU with the separate detectors.

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In the absence of ARU in the diagram the series/row of undesirable phenomena is observed. Under the influence of large signals amplifier instruments enter into the mode/conditions of a deep saturation. If as UP vacuum lamps are used, upon the saturation the following processes are observed: considerable circuitual current of control electrode of tube appears, in consequence of which the input resistance of cascade/stage descends and recovery time of the sensitivity of amplifier after the break-down of signal increases; the current of the screen grid of tube sharply grows, in consequence of which is reduced voltage/stress on the screen grid and, consequently, also is reduced the slope/transconductance of tube and the output voltage/stress of cascade/stage.

For the rapid restoration/reduction of the maximum sensitivity of amplifier after the action of large signals it is necessary that the handled cascades/stages would rapidly restore maximum sensitivity. For this it is necessary to provide the number of the measures, which ensure the virtually instantaneous discharge of transient capacity/capacitance after the break-down of large signal. If amplifier stage is assembled on the diagram with the series feed, it is necessary to apply the diagram of amplifier stage with parallel feed and choke/throttle in the circuit of control electrode of tube.

In order to attain the constancy of the output voltage/stress of cascade/stage in the saturation mode, it is necessary to ensure the constancy of voltage/stress on the screen grid of the tube: a) by the selection of the largest possible value of capacity/capacitance C_s of that shunting screen grid. During the amplification of pulse signals the time constant of capacity/capacitance C_s and anode resistance on the screen grid during the overloading must be large in comparison with the repetition period of impulses/momenta/pulses. In this case the voltage/stress on the screen grid remains in effect constant. The limiting factor are the overall sizes of capacitor/condenser; b) by application of voltage on the screen grid of tube from the voltage divider. The values of the

resistors/resistances of divider are chosen by such that the current of divider would be 10 times more than the current of screen grid. In this case capacity/capacitance C , it is possible to take somewhat smaller than in the first case; c) by application of voltage on the screen grid is direct from the anodic power supply through the choke/throttle. In this case capacity/capacitance C , can be selected smallest of all three cases examined. The constancy of voltage/stress on the screen grid is ensured by the constancy of the voltage of anodic power supply.

Vacuum-tube amplifier with ARU due to the nonlinearity of the characteristic of tube is to a considerable extent deprived of these deficiencies/lacks.

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If as UP transistors are used, then with the work of cascade/stage in the saturation mode base current sharply increases, in consequence of which the transistor can fail; increases recovery time of maximum sensitivity as a result of the fact that the transition of transistor from the saturation mode into the normal mode of work flows/occurs/lasts for a comparatively wide interval of time; parasitic oscillation [98] passes transistor into the converted mode/conditions and in the amplifier appears. The converted

mode/conditions appears on the high level of the signal, when the bias voltage of the transitions of transistor becomes opposite on the sign relative to the bias voltage normal mode.

In the presence in diagram of reverse mode it is sufficient one random impulse/momentum/pulse so that as a result of collision excitation of parasitic oscillations would arise. Physics of the onset of parasitic oscillations in the high-frequency transistors during the converted mode/conditions of transistor is described in work [98].

Parasitic oscillation with the overloading of transistor can be reduced, join up of collector/receptacle consecutively/serially with the oscillatory circuit (between the collector/receptacle and the duct/contour) active grid suppressor on the order of hundreds of ohms; increasing direct/constant voltage of the feed of the circuit of collector/receptacle and applying efficient diagram of ARU.

Consequently, in the amplifier it is expedient to apply ARU, based on the nonlinearity of the characteristics of transistor,. However, in the wideband amplifiers with fulfilling of inequality $\frac{U}{I} > 0,5$ this ARU cannot be applied. In this case the operational conditions of transistors must be chosen so that with the overloading of transistors the latter would not go out of order due to the

excessive increase of base current.

During the selection of the point of the connection of detector it is necessary to proceed from, in the first place, the input capacitance of diode in the smallest measure would affecting the resonance frequency of cascade/stage, which is especially important in the wideband amplifiers with small capacities/capacitances in the ducts/contours, and, in the second place, the signal, which enters the detector, must be sufficient to large ones so that the detector would work in the linear conditions. The conditions examined are contradictory, since, on the basis of the first condition, the detector must be connected partially to the load, and on the basis of the second - completely. Concrete/specific/actual circuit solution depends on the type of detector and required slope/transconductance of FAKh of amplifier.

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As the detectors the semiconductor diodes or transistors can be used. According to (4-1) the resistance/resistor of the load of detector R_n (Fig. 70) is chosen from the condition

$$R_n > 10R_{n.0} = 5p.$$

Transmission factor of the detected voltage/stress on the total

load

$$k = k_0 k_p = k_0 \frac{\rho}{2R_n + \rho} \quad (4-15)$$

Changing resistor/resistance R_n , it is possible to obtain the appropriate value of the coefficient of transmission k for each cascade/stage.

The possible versions of the connection of capacity/capacitance C_n , blocking resistor/resistance R_n in the high frequency, they are shown in Fig. 70.

The value of the time constant of detector $\tau_n = C_n R_n$ in narrow-band FU calculates as for the usual detectors, on the basis of the compromise condition of obtaining the minimum pulse rise-time and sufficient filtration of high frequency. For broadband FU without the delay line the time constant τ_n should be calculated from the condition of obtaining the constancy of the delay time of total output pulse.

The value of output voltage/stress, with which sets in the saturation of UK, to a considerable extent depends on the input resistance of the following cascade/stage. From this point of view all cascades/stages, with exception of the latter, work in the identical conditions. So that the level of the saturation of

latter/last cascade/stage on the output voltage/stress would not be higher than in the rest, its output sometimes it is expedient to shunt by the further semiconductor diode, which limits the output voltage/stress (Fig. 70).

The diagram of logarithmic band of UPCh on the transistors, given in Fig. 71, is an example of the practical realization of diagram with the separate detectors. In this diagram as the amplifier nonlinear cascades/stages the cascades/stages with ARU, based on the nonlinearity of the characteristics of transistors, are used. Amplifier stages are assembled on the common-base circuit on the high-frequency transistors $T_{11}-T_1$, of the type P417. Two-circuit band-pass filters with the capacitive coupling are load. The first duct/contour is parallel, is second - consecutive.

The second oscillatory circuit with the input of the following cascade/stage is connected autoinductive. For obtaining the wide passband and guarantee of stability of the characteristics of amplifier into the ducts/contours high attenuations are introduced, and, into the collector ducts/contours - in parallel (resistor/resistance $R_{11}-R_{11}$), and into the emitter ones - consecutively/serially (resistor/resistance $R_{11}-R_{11}$).

The selected diagram of ducts/contours is simple by the construction/design, is easily tuned and has good recurrence of the parameters.

Chains/networks of ARU are connected in the emitter circuit (resistor/resistance $R_1, -R_n$, and capacity/capacitance $C_1, -C_n$).

The emitter circuits of UP are fed from the separate source. Therefore the resistors/resistances of chains/networks of ARU $R_1, -R_n$, it is possible to select sufficiently large and to obtain required AKh of the cascades/stages (it is necessary to ensure inequality $I_{a.a.} R_1 \gg U_{a.a.}$), sufficient temperature and temporary/time [48] stability of the parameters of amplifier.

As the corrective elements/cells the separate transistor detectors (on transistors $T_1, -T_n$ of the type P416) are used, which have the following advantages over the diode ones: a comparatively high resistor/resistance, larger transmission factor, and with the help of them it is possible to carry out a good decoupling of UK in the high frequency.

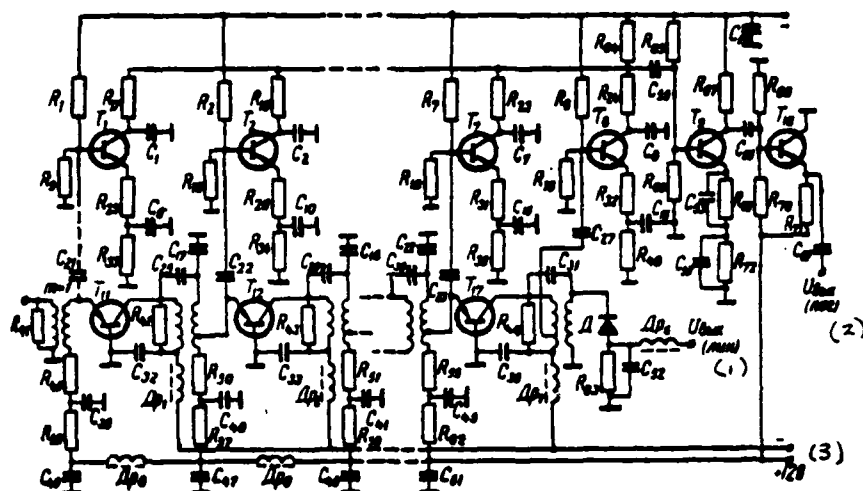


Fig. 71. The schematic diagram of logarithmic UPCh on the transistors with the separate detectors: T_1-T_5 - P416; $T_{11}-T_{15}$ - P417; R_1-R_{10} - 27 kilohm; $R_{11}-R_{15}$ - 2 kilohm; $R_{21}-R_{25}$ - 3.6 kilohm; $R_{26}-R_{30}$ - 30 ohms; $R_{31}-R_{40}$ - 2.7 kilohm; $R_{41}-R_{45}$ - 60 kilohm; $R_{46}-R_{50}$ - 2.7 ohms; R_{51} - 100 ohms; C_1-C_5 - 5 pF; $C_{11}-C_{15}$ - 150 pF; $C_{21}-C_{25}$, $C_{31}-C_{35}$, $C_{41}-C_{45}$ - 20 pF.

Key: (1). linear. (2). logarithmic. (3). +12 V.

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Usually, in the narrow-band logarithmic amplifiers as the summator delay lines with the lumped parameters are used. The diagram in question is broadband (passband of one cascade/stage $P \approx 12$ MHz) and group signal delay in the cascades/stages it is possible to disregard. Taking into account this fact, and also that the detectors ensure a good decoupling of cascades/stages in the high frequency, from the diagram is excluded the delay line and as the adder is used the resistor/resistance $R_1 = R_{\Sigma}$ which is the total load of all detectors. The considerable decrease of the distortions of the form of the detected video pulse is the corollary of this.

The inputs of detectors are connected to the ducts/contours autoinductive, in consequence of which they little affect the resonance frequency of amplifier.

For the linear addition of the output voltages/stresses of detectors must be performed the inequality

$$R_{\Sigma \text{ max. det}} > 10R_1.$$

(4-16)

For this with the collector circuits of detectors are connected the resistance/resistors of loads $R_n(R_{17}-R_{22})$, which are selected from the condition

$$R_n > 10R_z. \quad (4-17)$$

Together with capacitors/condensers C_1-C_4 , they fulfill the functions of the high-pass filters, which ensure the decoupling of amplifier stages. With the emitter circuits of the transistors of detection cascades/stages are connected control resistors $R_p(R_{23}-R_{32})$, which make it possible to change the transmission factors of detectors k_n and thereby to regulate accuracy and slope/transconductance of LAKh.

The accuracy of LAKh of multistage amplifier it is possible to obtain the higher, the less the scatter along the levels of limitation in nonlinear cascades/stages $u_{\text{lim}, n}$ (Fig. 17). Due to the scatter of the parameters of transistors, voltages/stresses $u_{\text{lim}, n}$ are different in different nonlinear cascades/stages. For increasing the accuracy of LAKh and the recurrence of the parameters of amplifier, it is desirable to make limitation level by the independent variable from the parameters of transistors. For this purpose in the diagram (Fig. 71) the mode/conditions of the work of transistors is selected so that the limitation in them begins somewhat earlier than in amplifier stages. In this case levels $u_{\text{lim}, n}$ are caused by limitation in the detectors, they depend in essence on the selected

modes/conditions of the work of transistors (supply voltage) and therefore they have a small scatter and it is sufficiently stable.

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The semicascade amplifier has the following parameters: at the work in linear conditions $K_v = 3.5 \cdot 10^4$; $P = 4$ MHz; $f_v = 30$ MHz; dynamic range of LAKh without the detector, connected at the input of amplifier, (T_1) $D_{MAX} \approx 82$ dB; $U_{n.n} \approx 2 \cdot 10^{-6}$ in; $\sigma = 120$ mv/dB; $\delta\sigma = 15\%$. Upon the start of null detector (T_T) the range increases to $D_{00} \approx 90_{00}$.

An amplitude characteristic (Fig. 107) and an example of calculation are given in Chapter 6.

If in the diagram the diode detectors, which work without the limitation, are used, in the UK it is necessary to ensure the constancy of the output voltage/stress of amplifier stages, which work in the mode/conditions of limitation. For obtaining by precise LAKh it is necessary that the extent of the horizontal section AKh of cascade/stage would be not less than the working dynamic range d. It is possible not to be interested in the form of the amplitude characteristic of cascade/stage with further increase of signal, since the voltage on the input will cease to grow due to the limitation in the previous cascade/stage.

The simplest and efficient method of obtaining the horizontal section is the inclusion into the collector circuit of effective resistance R_{or} (Fig. 54d). In this case with a sharp increase of collector current descends the supply voltage on the collector/receptacle of transistor and, consequently, also the level of limitation. Usually the value of resistor/resistance R_{or} lies/rests within the limits of 300-500 ohms. The amplitude characteristics of cascade/stage from ARU on transistor P417 for the different values of resistor/resistance of R_{or} and modes of feeding of transistor are given in Fig. 72.

During the amplification of radio pulses the capacity/capacitance of capacitor/condenser C_{or} blocking resistor/resistance R_{or} by high frequency, is chosen such value that its resistor/resistance would be sufficient to small ones for the reinforced frequencies, i.e., for variable component of collector current, and it is sufficient to large ones for slowly changing constant component of collector current.

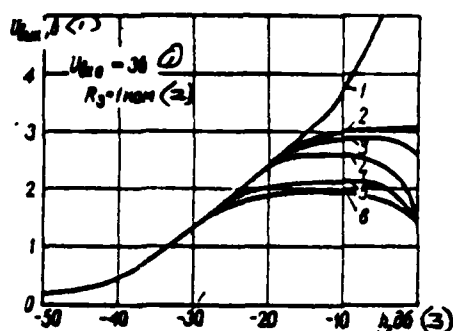


Fig. 72. The amplitude characteristics of cascade/stage with the limiting resistor/resistance: 1 - $E_N = -6$ B; $R_{or} = 0$; 2 - $E_N = -6$ B; $R_{or} = 300$ ohm; 3 - $E_N = -6$ B; $R_{or} = 390$ ohm; 4 - $E_N = -6$ B; $R_{or} = 470$ ohm; 5 - $E_N = -4$ B; $R_{or} = 390$ ohm; 6 - $E_N = -4$ B; $R_{or} = 470$ ohm.

Key: (1). B; (2). kilohm. (3). dB.

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However, capacitance value C_{or} must be such that the inertness of control on the voltage/stress on the collector/receptacle would be sufficient small and control was realized according to the law of the envelope of the radio pulse. Chain/network $R_{or}-C_{or}$ can be used for the suppression of prolonged interference.

The diverse variants of the schematics of logarithmic amplifiers are given in works [7, 8].

LOGARITHMIC SELECTIVE AMPLIFIERS WITH CATHODE AND EMITTER DETECTION.

Since the principle of construction of FU on the tubes and on the transistors is identical, it is expedient diagrams with the cathode (anodic) and emitter (collector) detection to consider together.

In the amplifier with the cathode detection the tubes of amplifier are placed in the mode/conditions of amplification and cathode detection, so that each of the cascades/stages, except voltage amplification with the oscillations of intermediate frequency and its supply to the following cascade/stage, detects the voltage/stress of radio pulses and gives independent of other cascades/stages the component of the output voltage/stress of video pulse on the total load.

The simplified circuit of two cascades/stages on the pentodes with the cathode detection is shown in Fig. 73. Sufficiently high resistor/resistance R_k in the cathode circuit of tube of UK is the resistance/resistor of the load of cathode detector. The diagram of cascade/stage with the cathode detection does not differ from diagram with ARU in pulse envelope.

The FAKh of multistage amplifier in this diagram is obtained just as in the diagram with the separate detectors. From the cathode resistor/resistance of each cascade/stage the detected voltages/stresses of the video pulses of positive polarity are removed/taken and they enter delay line, where they store/add up on total load $R_{\Sigma} = \frac{\rho}{2}$.

Amplitude characteristic of UK on the radio-voltage is the same as in the cascade/stage with ARU.

the simplest diagram with the cathode detection has the essential deficiencies/lacks, basic of which are:

- 1) the penetration of the voltage of video pulse because of this with the saturation of one of the cascades/stages the bias voltage in all amplifier stages automatically is changed.

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Therefore it is difficult to fit the necessary modes of operation of cascades/stages and to obtain LAKh of multistage amplifier in the broad dynamic band;

2) the presence of connection/communication between the cascades/stages on the video frequency, in consequence of which the amplifier works unstably and is inclined to the self-excitation.

For eliminating these deficiencies/lacks between the cathode of each tube of amplifier stage and the delay line it is possible to include/connect diode (Fig. 73), high resistor/resistance or separating cascade/stage. Diode must be switched on so that its resistor/resistance would be small for the passage of the current of video pulse, taken from the cathode of this tube, and large for the current of video pulse, which penetrates from other cascades/stages. If for the decoupling of cascades/stages vacuum or semiconductor diodes are used, then as a result of the low values of resistors/resistances ρ and resistors/resistances of diodes resistors/resistances R_n and cathode circuit of tubes are strongly shunted; therefore the magnitude of the detected video voltage and effect of control in the cascades/stages descends. Resistors/resistances R_n will not be shunted, if instead of the diodes as the elements/cells of decoupling to use resistors/resistances high in the value. In this case the voltage/stress of the video pulse, which enters the delay line, strongly decreases, i.e., the transmission factor of cascade/stage on

the video voltage considerably decreases. Of these deficiencies/lacks is deprived the diagram, in which as the elements/cells of decoupling are used corrective UK, with the help of which it is possible not only to completely untie the cathode circuits of cascades/stages in the high frequency, but also to correct the amplitude characteristic of an n-cascade amplifier.

The schematic diagram of the eight-stage logarithmic amplifier with the cathode detection and the corrective (untying) cascades/stages ($L_{1,0}$ - $L_{1,4}$) is given in Fig. 74.

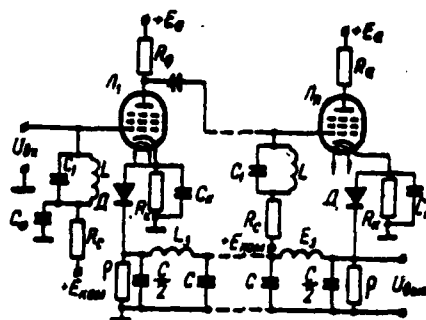


Fig. 73. The simplified circuit of logarithmic amplifier with cathode detection.

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Amplifier has the following parameters: $K_s=10^3$; $f_s=25$ MHz; $P=1$ MHz without the filter with the concentrated selection (FSS) and $P=0.5$ MHz with FSS; $D_{MAX} \approx 100$ dB; $U_{sx, u} \approx 5 \cdot 10^{-4}$ V; $\sigma \approx 0,22$ V/dB when $k_1 = 1$; $\delta\sigma=10-15\%$; $\delta U=2-3\%$.

With the cathode circuits of amplifier stages are connected the chains/networks of ARU ($R_1, R_2, R_3, \dots, C_1, C_2, C_3, \dots$), from which positive video pulse is removed/taken.

So that the grid currents of the tubes of amplifier stages would not affect the effect of adjustment and the form AKh of cascade/stage, the compensating voltage/stress in the circuit of the

grids of tubes is supplied from the source with the low internal resistor/resistance (L).

Amplifier has two outputs: linear with the limitation from the output of detector (left half L ,) and logarithmic from the delay line. Voltages/stresses from two outputs are summarized with the summator (L_1), which has two outputs: total U_{sum} and logarithmic U_{log} .

For obtaining LAKh of high accuracy the transmission factors of the corrective cascades/stages can be changed, changing the values of negative feedback on alternating current of cathode. This is achieved by a change in the resistors/resistances of feedback (not blocked parts of potentiometers R_{11} , R_{12} , R_{13} , and so forth). With the displacement/movement of the wipers the depth of OOS on alternating current is changed, mode/conditions on the direct current remains constant.

So that the untying cascades/stages would work stably, with the cathodes of tubes the resistors/resistances of feedback (R_{11} , R_{12} , ...) were connected. As the summator the delay line, connected with the anode circuits of the corrective cascades/stages, is used.

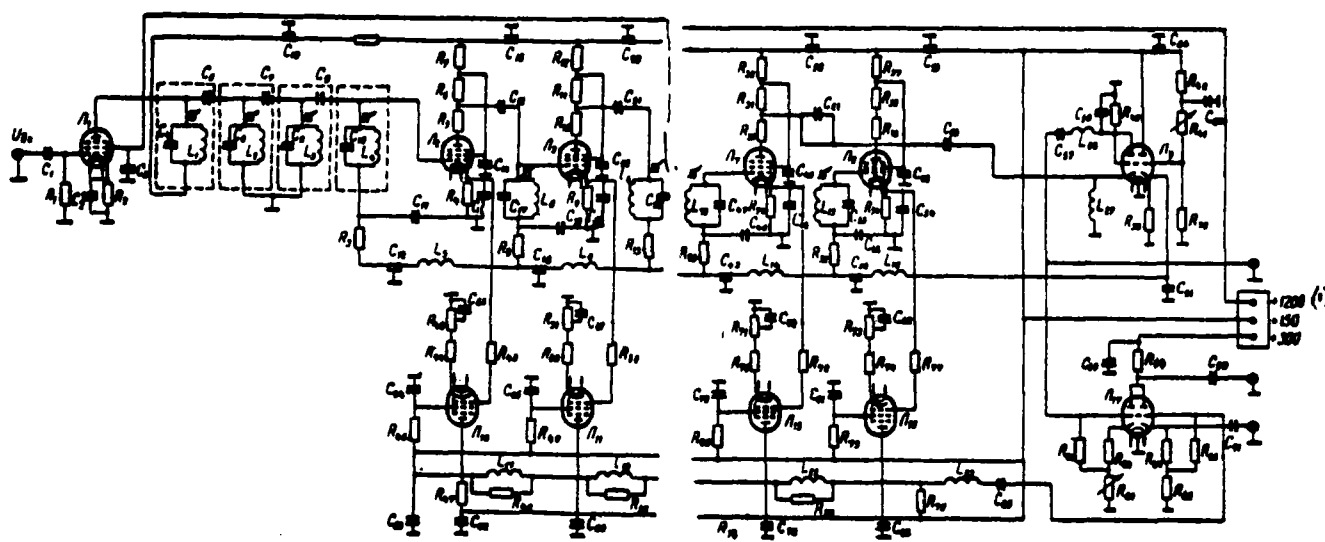


Fig. 74. The schematic diagram of logarithmic amplifier with the cathode detection and the corrective cascades/stages: L_1 - L_5 ; L_{11} - L_{14} - 6Zh1P; R_4 , R_5 , R_{14} , ..., R_{34} - 3.6 kilohm; R_{43} , R_{51} , R_{53} , ..., R_{71} - 1.1 kilohm; R_{44} , R_{50} , R_{54} , ..., R_{74} - 3 kilohm; R_5 , R_{10} , ..., R_{35} - 100 ohms; R_6 , R_{11} , R_{16} , ..., R_{56} - 2.7 kilohm; R_7 , R_{12} , ..., R_{37} - 100 Ohm; C_{13} , C_{23} , C_{30} , ..., C_{54} - 150 pF; C_{12} , C_{16} , C_{18} , C_{20} , ..., C_{55} - 3300 pF.

Key: (1). V.

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The advantages of the diagram: in the diagram it is possible to

obtain high accuracy of LAKh in ShDD; diagram has the high stability of the parameters (form of LAKh and amplification factor), caused by deep OOS on the direct current in amplifier stages and by OOS on alternating and direct current in the corrective cascades/stages, and also a good recurrence of the parameters from one copy to the next, in consequence of which it is possible to perform multichannel receivers.

The amplitude characteristics of amplifier at the output of detector (broken line) and delay line with different feeding voltages/stresses on the screen grid are given in Fig. 75. From the figure one can see that the mode/conditions of limitation at the output of detector and form of LAKh barely are broken with a change in the feeding voltages/stresses within considerable limits. The analogous stability of characteristics is observed with a change in the voltage/stress in the filament circuit of tubes.

For the realization of the broadband logarithmic amplifier with the cathode detection, intended for amplifying the narrow pulses $t < 0.1 \mu s$, it is necessary as the load of amplifier stages to use two-circuit band-pass filters; resonance frequency to choose order 60-80 MHz; instead of the delay line to use the broadband cascades/stages of the amplification of video pulses, which have linear amplitude characteristic in the ShDD.

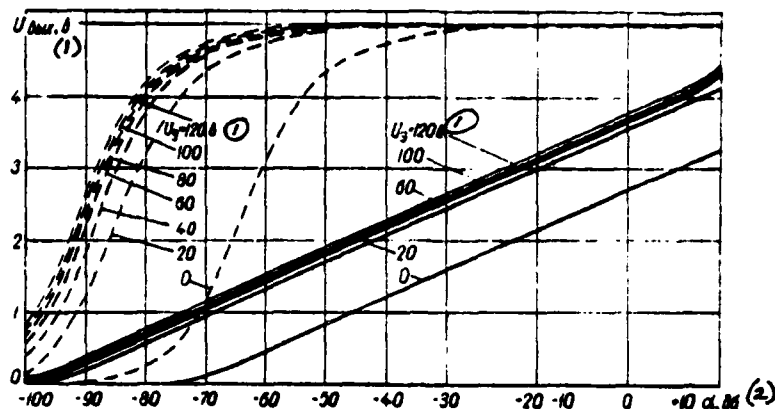


Fig. 75. Amplitude characteristics of amplifier (Fig. 74) with different voltages/stresses on the screen grids of the tubes: — on the video voltage; - - - on the radio voltage; $U_{\text{ex. 006}} = 0.5$ V; $E_a = 166$ V; $E_k = 20$ V; $R_{\text{o.o.}} = 3$ kilohm.

Key: (1). V. (2). dB.

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As a rule, in these cascades/stages must be realized high-frequency correction and deep negative feedback.

FAKh

The circuit solutions of amplifiers with ~~FAKh~~ by the method of addition can be realized much. The FU on the transistors it is possible to perform by analogy with vacuum-tube amplifiers. With some diagrams (with the grid, anodic and collector detection the

use/application of cascode diagrams) the reader can be introduced in works [7, 8]. In particular, work [8] gives the schematic diagrams of logarithmic transistorized amplifiers with the parameters:

seven-cascade transistorized amplifier P417 with the emitter detection without the untying cascades/stages: $K_v = 3.5 \cdot 10^4$; $P = 3.4$ MHz; $D = 80$ dB; $U_{\text{ex. n}} = 2 \cdot 10^{-6}$ V; $\sigma = 14$ mV/dB; $\delta\sigma = 20\%$; $f_c = 30$ MHz;

five-stage amplifier with the untying cascades/stages:
 $K_v = 3.1 \cdot 10^3$; $P = 4.1$ MHz; $f_c = 30$ MHz; $D = 70$ dB; $\sigma = 70$ mV/dB when $k_0 = 1$;
 $\delta U = (2-3)\%$; $\delta\sigma = 15\%$.

§ 4. Functional amplifiers on the parallel pairs.

The advantage of parallel pairs of FU is the fact that with their aid it is possible to design amplifier with FAKh when $b = \text{var}(\downarrow)$ in the ShDD on the amplifier instruments, which have a small dynamic range, in particular, on the transistors. Furthermore, there can be created FU intended for amplifying of both the video pulses and the radio pulses.

With the help of the parallel pairs the logarithmic amplifiers with the radio-output can be successfully created. The principle of parallel pairs can be used for the construction of FU on transistors

and on vacuum tubes.

APERIODIC LOGARITHMIC AMPLIFIERS OF VIDEO PULSES.

Parallel pair must consist of amplifier stage, which works of the mode/conditions of a linear-nonlinear amplification and limitation, and repeater, which works in the linear conditions. Expressions for the amplitude characteristics of UK and the repeater are given in Table 4.

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It is most easy to obtain the required characteristic of cascade/stage in the transistorized amplifier.

The mode/conditions of limitation in the transistor cascade/stage can occur due to the saturation of transistor or due to the cutoff. In both cases the value of the input and output voltages, with which the limitation begins, depends on the selected operating point and the value of load.

Let us consider saturation modes and cutoffs, which attack respectively during the amplification of negative and positive pulses in the transistor of type p-n-p.

Deficiencies/lacks in the transistor in the saturation mode:

1) the low input resistance, which even more is reduced with an increase in the signal;

2) the considerable expansion of the output pulse, whose value depends on the level of input pulse. This is explained by the fact that the process of saturation is accompanied by the delay of output signal relative to input, which is connected with the resorption of the surplus concentration of minority carriers in the base layer of the saturated transistor:

3) considerable recovery time of the maximum sensitivity of amplifier after the termination of the effect of strong signal;

4) a small dynamic range of AKh on the input voltage. For example, if we take operating point for the high-frequency transistor when $U_{a.e.} = -0,22$ V, then for the load resistance/resistors, used in the video amplifiers, saturation sets in on input level on the order of 70-100 mV.

If one considers that linear conditions of amplifier is finished

with $U_{\text{н.н.}} = 10-15$ mV, then for observing the equality $K_{\text{н.н.}} = d$, which must be performed with the successive work of cascades/stages, the coefficient of amplifier $K_{\text{н.н.}}$, must be order 6-7, which is clearly insufficient.

Advantages of cutoff conditions in comparison with the saturation mode:

1. Good coincidence of real AKh with that required, described by the expressions, given in Table 4.
2. Short recovery time of maximum sensitivity of amplifier after break-down of strong signal.

An increase in the input resistance of cascade/stage with an increase in the signal is a deficiency/lack in the cutoff conditions..

As we see, cutoff conditions, which should be applied in the logarithmic amplifiers, is best.

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The level of limitation (cutoff) can be varied, changing the position of operating point and the value of the load

resistance/resistor.

It is necessary to note that the real amplitude characteristic of the UK can coincide in form with the required characteristic only with certain approximation/approach. Ideal coincidence of characteristics cannot be obtained.

The schematic diagram of transistor logarithmic video amplifier on the parallel pairs is given in Fig. 76. Amplifier stages are made on transistors T_1, \dots, T_n , repeaters - on transistors T_1, \dots, T_n . For obtaining the considerable dynamic range of linear characteristic and required transmission factors the repeaters on the collector circuit are fed from the source -30 V and are made with variable/alternating/variable current feedback of signal in the emitter circuit (potentiometers R_1, \dots, R_n , depicted in Fig. 76).

For obtaining the precise LAKh n-cascade amplifier it is necessary to ensure the strictly successive work of identical nonlinear cascades/stages.

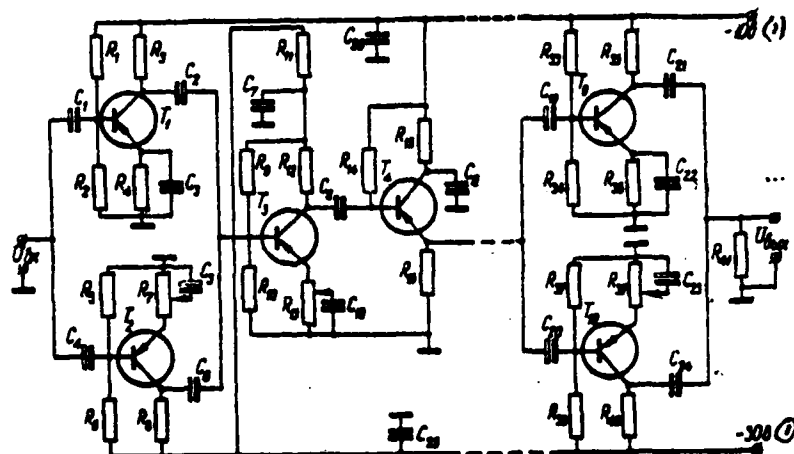


Fig. 76. The schematic diagram of logarithmic video amplifier on the parallel pairs: T_1, T_2, T_3 - P402; R_1, R_3 - 30 kilohm; R_2, R_5, R_3, R_4, R_7 - 6.8 kilohm; R_6, R_8 - 3.6 kilohm; R_4, R_5 - 1.8 kilohm; R_9, R_{10} - 24 kilohm; R_1, R_3 - 3.3 kilohm; R_6, R_{10} - 5.1 kilohm; R_{10}, R_{11} - $C_1, C_3, C_7, C_{10}, C_{21}, C_{23}, C_{25}, C_{26}$ - 20 μF .

Key: (1). V.

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With the direct connection of nonlinear cascades/stages to satisfy the condition for successive work is virtually impossible, since amplifier stage, assembled on the diagram with OE, turns over the phase of the reinforced signal by 180° .

If in the amplifier transistors of the type p-n-p are used only, then for the realization in each cascade/stage of cutoff conditions it is necessary that the input of each cascade/stage the pulse of signal only of positive polarity would enter. For this it is necessary in the multistage amplifier between the nonlinear cascades/stages to switch on the phase-inverting cascade/stage with the transmission factor, equal to one. This phase-inverting cascade/stage in the diagram (Fig. 76) is assembled on transistor T₁. The phase-inverting cascade/stage must have linear characteristic in the ShDD at the input signal with the transmission factor, close to one. For fulfilling these requirements, in the cascade/stage the adjustable deep negative feedback is used, and the circuit of collector/receptacle is fed from the source with an increased voltage/stress of 30 V.

The input resistance of UK with the increase of signal increases; therefore for guaranteeing the strictly successive work of the cascades/stages between the nonlinear cascades/stages, besides the phase-inverting cascade/stage, it is necessary to switch on the emitter follower with the large and the constant resistance. The product of the transmission factors of the phase-inverting cascade/stage and the emitter follower must be equal to one.

Three-stage amplifier has following data:

$$K_0 = 8 \cdot 10^3; D_{\text{MAX}} \approx 75 \frac{\text{dB}}{\text{dB}}; U_{\text{B.E.}} \approx 3 \cdot 10^{-4} \frac{\text{V}}{\text{V}}; \delta U \approx (3-4) \%$$

Key: (1). dB. (2). V.

During the use in the amplifier of transistors of the type the p-n-p and n-p-n phase-inverting cascades/stages they are not necessary. The schematic diagram of the two-stage amplifier, in which transistors of both types are used, is depicted in Fig. 77. In the diagram between the nonlinear cascades/stages the emitter follower is connected. The first cascade/stage is assembled on transistors T_1 and T_2 , type P402, the emitter follower and the second cascade/stage are assembled on transistors T_3 , T_4 and T_5 , type P503A.

For expanding the dynamic range of the linear section of the characteristics of repeaters, which form part of nonlinear cascades/stages, the latter are fed from two series-connected power supplies on 15 V.

As a result of the conducted experimental investigations it is established that the LAKh of the amplifier of that assembled on the diagram (Fig. 76) on the germanium transistors, is sufficiently stable to temperature of 40°C.

If amplifier is assembled on the silicon transistors, the logarithmic law of amplification is retained to temperature of 60-65°C. The accuracy of the LAKh is higher, the greater the number of nonlinear cascades/stages in the amplifier.

In the amplifier it is possible to obtain a small parasitic reverse/inverse overshoot, after including/connecting the phase-inverting cascades/stages and cutoff conditions on the parasitic overshoot, what is the advantage of the amplifier examined in comparison with others.

LOGARITHMIC AMPLIFIER WITH THE RADIO-OUTPUT.

The logarithmic amplifier on the pairs, intended for amplifying the harmonic oscillations, to create considerably more difficult than video amplifier. This is caused by the fact that the phase shifts in amplifier stages and repeaters must be equal and since voltage on the input UK repeater is one and the same, at the end of the dynamic range of LAKh this voltage/stress can exceed that permitted for the transistor of amplifier stage.

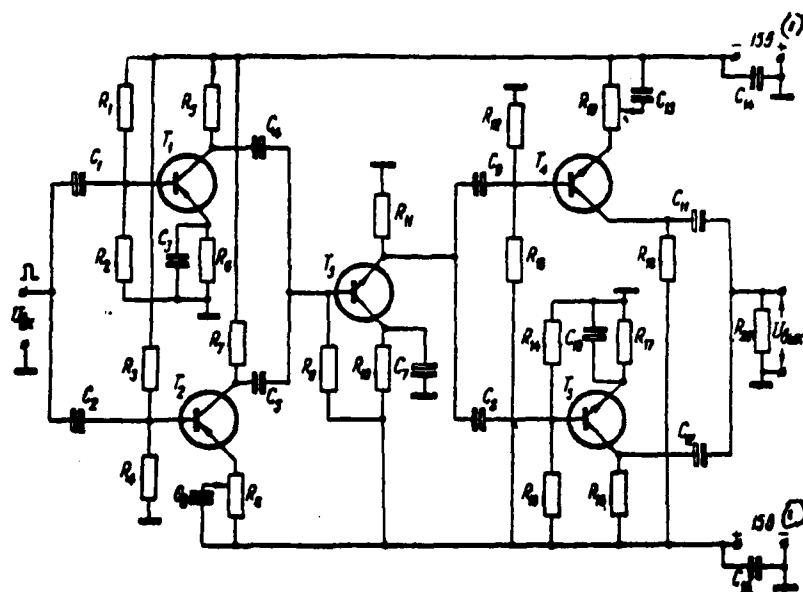


Fig. 77. Schematic diagram of the logarithmic video amplifier, assembled on the transistors of two types p-n-p and n-p-n.

Key: (1). in.

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So that the transistor would not malfunction, it must be protected by the limiter. Limitation level it is necessary to choose somewhat higher than the level, on which in the UK the limitation begins.

Nonlinear amplifier stages must have amplitude characteristics the same as in the video amplifiers. Furthermore, to them are

presented the further requirements: a small phase shift; the low level of limitation.

The first requirement does not need special explanation, the second - follows from the limitedness of the dynamic range of repeater. Therefore, the lower the level of limitation U_{orp} , the less the stringent requirements are imposed on repeater.

The protection of cascades/stages from the large voltages/stresses can be made in the form of limiters of any type, which allow/assume at their input of the voltage/stress of the order of several volts (for example, consecutive type diode attenuators). A deficiency/lack in the diode limiters is the dependence of their input resistance on signal level and, therefore, considerable effect on the previous cascades/stages.

The limiter circuit, made in the form of the emitter follower (T_1 in Fig. 78), is more acceptable, the transmission factor of which with the large signals sharply is reduced, and transistor itself works with the voltages/stresses permissible for it.

As the repeater it is possible to use cascades/stages with the divider at the input (Fig. 38a) or amplifiers with a deep negative feedback (Fig. 38c). However, from the point of view of obtaining large dynamic range at frequencies $f \gg f_0$ the first diagram much better.

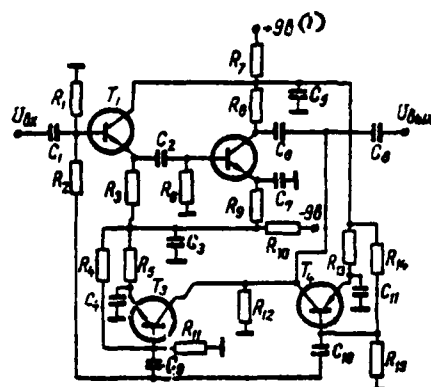


Fig. 78. The schematic diagram of the pair of the logarithmic amplifier of the harmonic oscillations: T_1, T_2, T_3 - P418Zh, T_4 - GT313B; R_1, R_6 - 1.3 kilohm; R_2, R_{12} - 680 ohms; R_3 - 2.4 kilohm; R_4, R_7 - 2.7 kilohm; R_5 - 62 ohms; R_{10} - 82; R_8 - 820; R_9 - 82; $C_1, C_2, C_3, C_4, C_5, C_6, C_7, C_8, C_9, C_{10}$ - 1 t pF; C_{11} - 10 t pF; C_{12} - 440 pF; C_{13} - 100 pF.

Key: (1). V.

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This is explained by the fact, in the first place, that at the high frequencies due to R_6 the complex connection/communication is created. It is possible to show that with the identical transmission factors $K_0 = 1$ the input resistance of diagram in Fig. 38a is more than in the diagram in Fig. 38c.

Furthermore, with the sinusoidal input voltage in the diagram on Fig. 38c voltage/stress $\dot{U}_{\text{в.с}}$ is also sinusoidal, and in the diagram in Fig. 38a when $R_0 \gg R_{\text{вх}}$ harmonic current I_0 . Since dependence $I = f(I_0)$ is virtually linear, and $I = f(U_{\text{в.с}})$ - is exponential, diagram in Fig. 38c is linear to the large input voltage.

The considerable phase shift at the high frequencies, which appears due to resistor/resistance R_0 is a deficiency/lack in the diagram (Fig. 38a). This deficiency/lack can be reduced by the start of capacitor/condenser C_0 in parallel to resistor/resistance. In this case from chain/network $R_0 C_0$ and $R_{\text{вх}} C_{\text{вх}}$ it is possible to form the frequency-independent divider.

The equivalent diagram of input circuit of transistor and chain/network $R_0 C_0$ is represented in Fig. 79.

The transmission factor of the repeater

$$K_{\text{с}} = |\dot{K}_{\text{дел}}| \cdot |K_1|,$$

where $|\dot{K}_{\text{дел}}| = \frac{|\dot{U}_{\text{вых. дел}}|}{|\dot{U}_{\text{вх. дел}}|}$ - transmission factor of divider;

$$|K_1| = \frac{|\dot{U}_{\text{вых}}|}{|\dot{U}_{\text{вх}}|} = \frac{|\dot{U}_{\text{вых}}|}{|\dot{U}_{\text{вых. дел}}|} - \text{transmission factor of cascade/stage.}$$

The transmission factor of repeater K_n is equal to one with satisfaction of the condition

$$k_{\text{rep}} = \frac{1}{K_1}. \quad (4-18)$$

For the transmission factor k_{rep} of diagram in Fig. 79 it is possible to register

$$k_{\text{rep}} = \frac{Y_1}{Y_1 + Y_2} = \frac{Y_1}{Y_2}, \quad (4-19)$$

where

$$Y_1 = \frac{1}{R_0} + j\omega C_0 = g_1 + jb_1; \quad Y_2 = \frac{1}{R_{\text{ex}}} + j\omega C_{\text{ex}} = g_2 + jb_2;$$

$$Y_2 = Y_1 + Y_3 = g_2 + jb_2$$

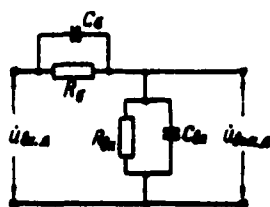


Fig. 79. Equivalent schematic of the frequency-independent voltage divider.

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We determine the modulus/module of transmission factor

$$k_{max} = \sqrt{\frac{(g_1 g_2 + b_1 b_2)^2 + (b_1 K_2 - g_1 b_2)^2}{g_1^2 + b_1^2}}. \quad (4-20)$$

from formula (4-19).

For determining the maximum value k_{max} we differentiate expression (4-20) on ω and let us make result equal to zero. After conversions we obtain

$$(g_2 C_0 - g_1 C_2)(\omega^2 C_1^2 + g_2^2) g_2 C_0 = 0.$$

Since

$$(\omega^2 C_1^2 + g_2^2) g_2 C_0 \neq 0,$$

that

$$g_2 C_0 = g_1 C_2, \quad (4-21)$$

or

$$g_1 C_{22} = C_0 g_2, \quad (4-22)$$

where

$$C_2 = C_0 + C_{\text{ex}}.$$

Into expression (4-21) the frequency does not enter. This means that k_{max} (for the linear circuit) is maximum at all frequencies, i.e., the divider is frequency-independent. Substituting expression (4-21) in (4-20), we obtain

$$k_{\text{max}} = \frac{a_1 a_2 + b_1 b_2}{a_2^2 + b_2^2}. \quad (4-23)$$

Solving together equations (4-21), (4-22) and (4-23), we obtain

$$R_0 = R_{\text{ex}} \frac{1 - k_{\text{max}}}{k_{\text{max}}}; \quad (4-24)$$

$$C_0 = C_{\text{ex}} \frac{k_{\text{max}}}{1 - k_{\text{max}}}, \quad (4-25)$$

where k_{max} is determined from expression (4-18).

In usual amplifier stage on the transistor it is difficult to obtain the linear section of amplitude characteristic higher than 1 V (taking into account that the input resistance of the following cascade/stage is its load). Thus, at the frequency of 30 MHz for the transistor of the type P418Zh with $I_n = 10$ mA and $U_n = 10$ into linear section $U_{\text{max}} = 0,8$ in, which is completely insufficient.

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For expanding the linear section it is possible to use the

series connection of transistors (Fig. 41b) with the simultaneous voltage doubling of feed. However, in this case with the large amplitudes limitation on the current begins. Most acceptable is the push-pull circuit, assembled on transistors T, of the type P418Zh (n-p-n- conductivity) and T, type GT313B (p-n-p- conductivity).

The schematic diagram of the parallel pair, assembled on the transistors of types P418Zh and GT313B, is depicted in Fig. 78. Amplifier stage is assembled on transistor T, of the type P418Zh. Selectivity in amplifier stage is realized by the frequency-dependent feedback on the emitter circuit.

The amplifier, which consists of the pairs, whose diagram is given in Fig. 78, and one amplifier stage on transistor P418Zh, connected at the input of amplifier, has the following parameters: $K_v = 1,8 \cdot 10^4$; $P = 10$ MHz; $f_c = 30$ MHz; $D_{\text{MAX}} = 80$ db; $U_{\text{max}} \approx 10^{-6}$ V; $\sigma = 22$ mV/dB; $\delta\sigma = 15-20\%$.

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Chapter 5.

TRANSIENT PROCESSES IN FUNCTIONAL AMPLIFIERS WITH THE NONLINEAR
ELEMENTS.

§ 1. Methods of the study of transient processes in the functional
amplifiers.

Any functional amplifier (with exception of linear) is nonlinear, even if it consists of linear cascades/stages. Because of this the transient processes, which take place in FU, have a number of special features/peculiarities. Such special features/peculiarities include a change in the form of front (in the region short times) and flat/plane part (in the region long times) of transient response with the increase of input signal level. The considerable distortions of the form of pulse signal with a change in its level are the corollary of this.

During the amplification of video pulses at the output of amplifier with $b = \text{var } (\downarrow)$ are formed the considerable parasitic reverse/inverse overshoots, which impede the practical use of functional video amplifiers with the dynamic range of more than 60 dB.

Furthermore, increases decay in the flat/plane apex/vertex of video pulse at the end at the end of the range of FAKh with $b = \text{var } (\downarrow, \uparrow)$, in consequence of which real FAKh additionally differs from precise.

With a change in the form of pulse edge and, consequently, also the time lag of impulse/momentum/pulse at the output of amplifier with an increase in the signal appears further of error in the information, placed in the leading impulse front of a small level.

For a study of transient processes in functional amplifier stage it is most expedient to use the method, which depends on the internal essence of the diagram of cascade/stage and form of the influencing effect (signal). In turn, the internal essence of diagram is determined by type of UP (linear or nonlinear, inertia-free or inertial) and by type of load, which is, as a rule, is inertial and can be linear or nonlinear.

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Possible combinations of UP and loads are shown in Fig. 80. Depending on diagram, one or another the method of study is applied. Let us consider the possible methods of the study of transient processes and the aperiodic nonlinear cascades/stages, on which the jump of direct/constant voltage or current acts.

SOLUTION OF PROBLEMS WITH THE HELP OF NONLINEAR DIFFERENTIAL EQUATIONS.

The possibility of the solution of problem in general is the advantage of the method of differential equations. If nonlinear differential equation is obtained with the divided variables, then problem is solved easily and rapidly. In the majority of the cases of compound circuits the nonlinear differential equations are fairly complicated, to solve which difficult, since the regular methods of solving the nonlinear differential equations in mathematics are not worked out.

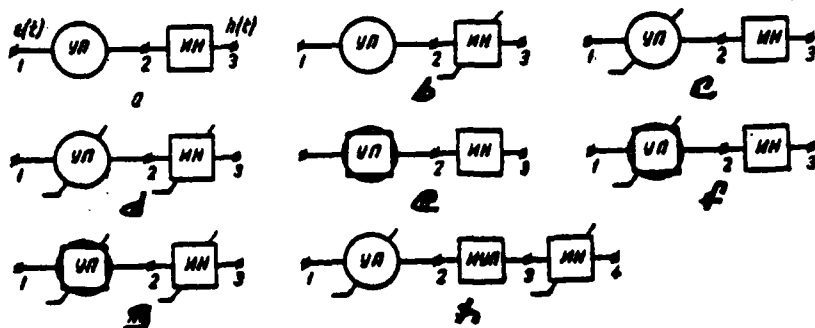


Fig. 80. The possible versions of the combination of amplifier instrument and load: UP, UP - linear and nonlinear inertia-free UP; UP, UP - linear and nonlinear inertial UP; NN, NN - linear and nonlinear inertial load; NN - element/cell, which considers inertness of UP.

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For the compound circuits nonlinear equations in the majority of the cases are not reduced to the equations with the divided variables and in a precise form are not integrated, in connection with which it is necessary to apply the approximation methods of integration.

Transient processes are described by equations of the Bernoulli type in the extremely limited number of diagrams, and the equation of types Riccati and Abel are expressed as quadratures in a few special cases.

However, research of the transient processes in the region of short and long times for the diagrams (Fig. 80) frequently is reduced to the research of transient processes in the simplest equivalent nonlinear diagrams, given in Fig. 81. Under the influence at the the input of the cascade/stage of the jump of direct/constant voltage or current, the nonlinear differential equation, which describes transient processes in the diagrams (Fig. 81), it is possible to register in the following form:

$$W\left(\frac{d}{dt}\right)x + f[x(t)] = N_m\varphi(t), \quad (5-1)$$

or

$$W\frac{dx}{dt} + f[x(t)] = n(t),$$

where X - unknown value (voltages/stresses or currents); $W(d/dt)$ - the temporary/time immittance, expression for which is defined by the classification of the linear elements of network (resistors/resistances, capacities/capacitances, inductance); N_m - amplitude of current or voltage of the equivalent generator, which substitutes UP; $\varphi(t)$ - function, which characterizes inertness UP. For inertia-free linear and nonlinear UP function $\varphi(t)=1$.

The integration of equation (5-1) in a precise form depends on the character of the right side of this equation $n(t)$. When $\varphi(t)=1$ the variables in equation (5-1) are divided:

$$dt = \frac{W dx}{N_m - f(x(t))}.$$

Integrating in the limits from 0 to t, we obtain

$$t = W \int_{x_0}^x \frac{dx}{N_m - f(x)}, \quad (5.2)$$

where x_0 - value of the unknown quantity with $t=0$, determined from the initial conditions.

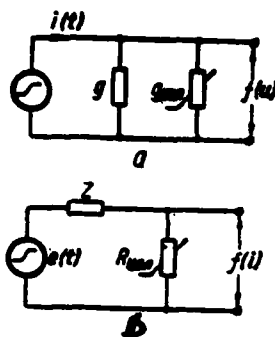


Fig. 81. Equivalent schematics of the simplest nonlinear systems with the equivalent current generators (a) and voltage/stress (b).

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If function $n(t)$ depends on time, variables in equation (5-1) in the general case are not divided and can be solved only approximately. It is most expedient to seek by the following methods:

to replace the real function $n(t)$ with linear segments $b_k t$ in the individual sections of time from t_{k-1} to t_k and to use the method of fitting;

to use expansion in the series/row.

Method, based on the use/application of Laplace transform.

The basis of method is the straight line

$$F(p) = Lf(t) = \int_0^{\infty} e^{-pt} f(t) dt \quad (5-3)$$

and reverse/inverse

$$f(t) = L^{-1}F(p) = \frac{1}{2\pi j} \int_{\sigma-j\infty}^{\sigma+j\infty} e^{pt} F(p) dp \quad (5-4)$$

of the Laplace transform. The use/application of direct transformation of Laplace (5-3) leads to the algebraization of differential equations and thereby facilitates the solution of some problems.

Applying to equation (5-1) the Laplace transform, we obtain

$$W(p)X(p) + F(p) = N(p).$$

Whence

$$X(p) = \frac{N(p)}{W(p)} - \frac{F(p)}{W(p)}, \quad (5-5)$$

where

$$X(p) = Lx(t); \quad W(p) = LW\left(\frac{d}{dt}\right); \quad F(p) = Lf[x(t)]; \\ N(p) = Ln(t).$$

After introducing the designation

$$H(p) = \frac{p}{W(p)}, \quad (5-6)$$

expression (5-5) can be registered

$$X(p) = \frac{N(p)H(p)}{p} - \frac{F(p)H(p)}{p}. \quad (5-7)$$

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Inverse transformation

$$h(t) = L^{-1}H(p). \quad (5-8)$$

is the original of image (5-6).

The unknown value $x(t)$ is inverse transformation of Laplace expression (5-7)

$$x(t) = L^{-1} \frac{N(p)H(p)}{p} - L^{-1} \frac{F(p)H(p)}{p}.$$

Using a theorem of superposition from the theory of the Laplace transform, we obtain expression for the original of function $x(t)$ under the zero initial conditions [$x(0)=0$ with $t=0$].

$$x(t) = \int_0^t h(t-\tau) n(\tau) d\tau - \int_0^t h(t-\tau) f[x(\tau)] d\tau. \quad (5-9)$$

If nonlinear element/cell is absent $f[x(\tau)] = 0$, expression (5-9) takes the form

$$x_0(t) = \int_0^t h(t-\tau) n(\tau) d\tau. \quad (5-10)$$

Then expression (5-9) can be registered

$$x(t) = x_0(t) - \int_0^t h(t-\tau) f[x(\tau)] d\tau. \quad (5-11)$$

The solution of integral (5-11) accordingly [95] are the recursion formulas

$$\left. \begin{aligned} x_0(t) &= \int_0^t h(t-\tau) n(\tau) d\tau; \\ x_1(t) &= x_0(t) - \int_0^t h(t-\tau) / [x_0(\tau)] d\tau; \\ x_2(t) &= x_0(t) - \int_0^t h(t-\tau) / [(x_1(\tau))] d\tau; \\ &\vdots \\ x_{n+1}(t) &= x_0(t) - \int_0^t h(t-\tau) / [x_n(\tau)] d\tau. \end{aligned} \right\} \quad (5-12)$$

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On the absolute convergence of series/row (5-12) it is proved in work [95]. With the help of integral (5-12) it is possible to solve series of problems in the transient processes in the nonlinear systems. A deficiency/lack in the method examined is the need of solving the large number of integrals. Problem is facilitated, if we use tables with the operational images. For this series/row (5-12) must be represented in the form of the series/row of operational images.

Using inverse transformation of Laplace for expression (5-5), we obtain

$$x(t) = x_0(t) - L^{-1} \left\{ \frac{Lf[x(t)]}{W(\rho)} \right\}, \quad (5-13)$$

where

$$x_0(t) = L^{-1} \frac{N(p)}{W(p)}.$$

Then series/row (5-12) can be registered in the following form:

$$\left. \begin{aligned} x_1(t) &= x_0(t) - L^{-1} \left\{ \frac{Lf[x_0(t)]}{W(p)} \right\}; \\ &\vdots \\ x_{n+1}(t) &= x_0(t) - L^{-1} \left\{ \frac{Lf[x_n(t)]}{W(p)} \right\}. \end{aligned} \right\} \quad (5-14)$$

The quantity of terms in series/rows (5-12) and (5-14), which must be taken during the calculation, is determined by the degree of the convergence of these series/rows and required accuracy of calculation. In the theory of iteration method it is proven [3], that the process of calculations according to the algebraic equations of form $x_i = f_i(x_1, x_2, \dots, x_n)$, ($i=1, 2, \dots, n$) converges with satisfaction of the conditions

$$\sum_{i=1}^n \left(\frac{\partial f_i}{\partial x_1} \right) < 1; \quad \sum_{i=1}^n \left(\frac{\partial f_i}{\partial x_2} \right) < 1; \quad \dots; \quad \sum_{i=1}^n \left(\frac{\partial f_i}{\partial x_n} \right) < 1. \quad (5-15)$$

The convergence the better, the the left sides of expressions (5-15) in comparison with one are less.

Thus, in each individual case the convergence of calculations

can be easily established/installed.

The presence in integrals (5-3) and (5-4) infinite limits indicates that the exact solution of series/rows (5-12) and (5-14) will be obtained with the striving count terms to infinity, i.e.

$$x(t) = \lim_{n \rightarrow \infty} x_n(t). \quad (5-16)$$

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With the slow convergence of series and the infinite limits hinders the calculation even in the contemporary computers, what is a deficiency/lack in the method examined.

This deficiency/lack in the operational method can be excluded, after fitting the integral transform, similar according to its structure to (5-3), but requiring integration within final limits. It is obvious, there is a series/row of conversions, which satisfy the conditions indicated. In particular, by G. Ye. Pukhov [55] as direct transformation it is proposed to use known formula [60]

$$\tilde{F}_v = j \frac{2}{T} \int_0^T e^{-jv\omega t} f(t) dt, \quad (v = 0, 1, 2, \dots, \infty), \quad (5-17)$$

making it possible to determine complex amplitude \tilde{F}_v of the v th order of function $f(t)$, assigned in the final segment $(0, T)$. Inverse transformation consists of the simple operation of the composition of

trigonometric Fourier series in known values F_n , i.e.

$$f(t) = \frac{1}{T} \sum_{n=-\infty}^{+\infty} e^{-jn\omega t} \int_0^T e^{-jn\omega t} f(t) dt. \quad (5-18)$$

If segment $(0, T)$ is replaced by $(0, 2\pi)$, then

$$\dot{F}_n = \frac{1}{\pi} \int_0^{2\pi} e^{-jn\theta} f\left(\frac{\theta}{\omega}\right) d\theta; \quad (5-19)$$

$$f(t) = \frac{1}{2\pi} \sum_{n=-\infty}^{+\infty} e^{jn\omega t} \int_0^{2\pi} e^{-jn\theta} f\left(\frac{\theta}{\omega}\right) d\theta, \quad (5-20)$$

where.

$$\theta = \omega t = 2\pi \frac{t}{T}; \quad t = \frac{\theta}{\omega} = T \frac{\theta}{2\pi}.$$

Analogous with the Laplace transform complex function \dot{F}_n is named the image of function $f(t)$, and function the real variable $f(t)$ - by original.

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For the brevity expression (5-17) sometimes is written/recorded in the following form:

$$\dot{F}_n = K_n [f(t)] = K_n f, \quad (5-21)$$

and straight/direct and inverse transformations:

during the operational calculus

$$f(t) \rightleftharpoons F(p); \quad F(p) \rightleftharpoons f(t); \quad (5-22)$$

during the complex calculation/enumeration

$$i(t) = \dot{P}_i; \quad \dot{P}_i = i(t), \quad (5-23)$$

§ 2. Transient processes in the aperiodic functional amplifiers with the nonlinear load.

Single-stage amplifier.

The equivalent diagram of nonlinear cascade/stage is depicted in Fig. 82a, in which amplifier instrument in the general case is substituted by inertial equivalent current generator $i(t) = S(t)V_{in}$. For inertia-free UP we have $i(t) = I_m$.

Transient processes we will examine on the assumption that at the moment of time $t=0$ at the input of nonlinear cascade/stage voltage surge U_{in} acts. Since the transient processes in the region of short times depend on the network elements, which are determining the frequency characteristic of the cascade/stage in the region of the highest frequencies, and the transient processes in the region of long times - by network elements, which are determining the frequency characteristic in the region of the lowest frequencies, it is expedient from the overall equivalent diagram of cascade/stage to isolate equivalent diagrams for the highest and lowest frequencies and to consider separately transient processes in each of these diagrams.

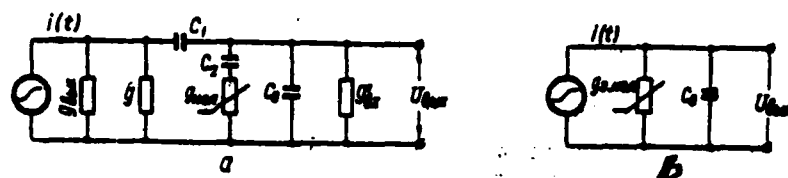


Fig. 82. Equivalent the diagram of the nonlinear cascade/stage: a) complete; b) for the region of the highest frequencies;

$g_{\text{nonl}} = g + g_{\text{nonl1}} + g_{\text{nonl2}} + g_{\text{nonl3}}$; g - conductivity of anodic or collector resistors/resistances.

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Transient processes in the region of short times. (Distortions of pulse edge).

The equivalent diagram of the cascade/stage for the region of the highest frequencies is depicted in Fig. 82b.

Let us consider transient processes in the amplifier with inertia-free UP (with a tube).

Inertia-free UP.

Under the influence at the input of nonlinear cascade/stage at the moment of time $t=0$ of voltage surge U_{in} transient processes in

the diagram are described by the nonlinear differential equation

$$\frac{\partial U_{\text{BHX}}}{\partial t} + \frac{g_{\text{B, нел}}}{C_0} U_{\text{BHX}} = \frac{S U_{\text{BX}}}{C_0}. \quad (5-24)$$

For guaranteeing the strictly successive work of nonlinear cascades/stages in a n-cascade amplifier amplifier the equivalent conductivity

$$g_{\text{B, нел}} = g_0 \varphi(z), \quad (5-25)$$

where the function $\varphi(z)$ is described by expressions given in Table 5.

$$\text{Conductivity } g_0 = g + g_{\text{BHX}} + g_{\text{BX}} + g_0 = \frac{1}{R_0}.$$

After substitution in equation (5-24) of conductivity $g_{\text{B, нел}}$, relative time α_0 and standardized/normalized voltages/stresses x and z we obtain

$$\frac{\partial x}{\partial \alpha_0} + \varphi(z) z = x(\alpha_0), \quad (5-26)$$

where

$$\alpha_0 = \frac{t}{C_0 R_0} = \frac{t}{\tau}.$$

Under the influence of unit function the variables in equation (5-26) easily are separated/liberated

$$\alpha = \int \frac{dx}{x(\alpha) - \varphi(z) z} + C. \quad (5-27)$$

Further it is necessary to compute integral (5-27) with

different functions $\phi(z)$.

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Linear conditions for all functional amplifiers. In this case $\phi(z)=1$ ($z \leq 1$). Result of calculating integral (5-27) taking into account the initial conditions

$$a_1 = \ln \frac{z}{z-1}. \quad (5-28)$$

Functional mode/conditions. Lowering cumbersome mathematical calculations, let us give the final results of calculating integral (5-27) taking into account the condition of the mating of linear and nonlinear sections of AKh for the most widely used types of FU.

The logarithmic amplifier with any foundation N

$$a_{11} = \frac{z-1}{z} + \ln \frac{z}{z-1} + \frac{a}{z} \ln \frac{z-1}{z-1-\frac{a}{z}}. \quad (5-29)$$

Exponential amplifier with $\beta = (1/n < 1)$, where $n=2, 3, 4, \dots$

$$a_{11} = \int \frac{dz}{z(a) - \left(\frac{z-1+n}{n}\right)^n} + C. \quad (5-30)$$

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The solution of integral (5-30) falls into two integrals ϵ'_{II} - for the n-positive, the even;

ϵ''_{II} - for the n-positive, the odd,

$$\begin{aligned} \epsilon'_{II} = x^{\frac{1-n}{n}} & \left[\ln \frac{1 + \frac{s-1+n}{\frac{1}{x^n}}}{1 - \frac{s-1+n}{\frac{1}{x^n}}} - 2 \sum_{k=1}^{\frac{n}{2}-1} p_n \cos \frac{2k}{n} \pi + \right. \\ & + 2 \sum_{k=1}^{\frac{n}{2}-1} q_n \sin \frac{2k}{n} \pi - \ln \frac{x^{\frac{1}{n}}}{1 - \frac{1}{x^{\frac{1}{n}}}} + 2 \sum_{k=1}^{\frac{n}{2}-1} p'_n \cos \frac{2k}{n} \pi - \\ & \left. - 2 \sum_{k=1}^{\frac{n}{2}-1} q'_n \sin \frac{2k}{n} \pi \right] + \ln \frac{x}{s-1}, \quad (5-31) \end{aligned}$$

where

$$\begin{aligned} p_n &= \frac{1}{2} \ln \left[\frac{(s-1+n)^2}{x^{\frac{1}{n}} n^2} - 2 \frac{s-1+n}{x^{\frac{1}{n}} n} \cos \frac{2k}{n} \pi + 1 \right]; \\ q_n &= \operatorname{arctg} \frac{\frac{1}{x^{\frac{1}{n}}} - \cos \frac{2k}{n} \pi}{\sin \frac{2k}{n} \pi}; \\ p'_n &= \frac{1}{2} \ln \left[\frac{1}{x^{\frac{1}{n}}} - 2 \frac{1}{x^{\frac{1}{n}}} \cos \frac{2k}{n} \pi + 1 \right]; \\ q'_n &= \operatorname{arctg} \frac{\frac{1}{x^{\frac{1}{n}}} - \cos \frac{2k}{n} \pi}{\sin \frac{2k}{n} \pi}; \quad h = s-1+n; \end{aligned}$$

$$a_n^* = \ln \frac{x}{x-1} + x^{\frac{1-n}{n}} \left[\ln \frac{x^{\frac{1}{n}}}{1 - \frac{1}{x^{\frac{1}{n}}}} + 2 \sum_{k=0}^{\frac{n-1}{2}} p_n \cos \frac{2k+1}{n} \pi + \right. \\ \left. + 2 \sum_{k=0}^{\frac{n-1}{2}} q_n \sin \frac{2k+1}{n} \pi - 2 \sum_{k=0}^{\frac{n-3}{2}} p_n^* \cos \frac{2k+1}{n} \pi - 2 \sum_{k=0}^{\frac{n-3}{2}} q_n^* \sin \frac{2k+1}{n} \pi \right], \quad (5-32)$$

where

$$p_n = \frac{1}{2} \ln \left[\frac{\frac{h^2}{2}}{x^{\frac{1}{n}} n^2} + 2 \frac{\frac{h}{1}}{x^{\frac{1}{n}} n} \cos \frac{2k+1}{n} \pi + 1 \right];$$

$$q_n = \operatorname{arctg} \frac{\frac{\frac{h}{1}}{x^{\frac{1}{n}} n} + \cos \frac{2k+1}{n} \pi}{\sin \frac{2k+1}{n} \pi};$$

$$p_n^* = \frac{1}{2} \ln \left[\frac{\frac{1}{2}}{x^{\frac{1}{n}}} + 2 \frac{\frac{1}{1}}{x^{\frac{1}{n}}} \cos \frac{2k+1}{n} \pi + 1 \right];$$

$$q_n^* = \operatorname{arctg} \frac{\frac{\frac{1}{1}}{x^{\frac{1}{n}}} + \cos \frac{2k+1}{n} \pi}{\sin \frac{2k+1}{n} \pi}.$$

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Exponential amplifier with $p=m>1$, where $m=2, 3, 4, \dots$

$$a_{II} = \int \frac{ds}{s(a) - (ms - m + 1)^{\frac{1}{m}}} + C. \quad (5-33)$$

The solution of integral (5-33) falls into two integrals: a'_{II} — for the index $1/m$ of positive, even

$$\begin{aligned} a'_{II} = \ln \frac{s}{s-1} + \left[\ln \frac{1 + \frac{\zeta}{s^m}}{1 - \frac{\zeta}{s^m}} - 2 \sum_{k=1}^{\frac{1}{2m}-1} p_k \cos 2km\pi + \right. \\ \left. + 2 \sum_{k=1}^{\frac{1}{2m}-1} q_k \sin 2km\pi - \ln \frac{1 + \frac{1}{s^m}}{1 - \frac{1}{s^m}} + 2 \sum_{k=1}^{\frac{1}{2m}-1} p'_k \cos 2km\pi - \right. \\ \left. - 2 \sum_{k=1}^{\frac{1}{2m}-1} q'_k \sin 2km\pi \right] x^{m-1}, \quad (5-34) \end{aligned}$$

where

$$\begin{aligned} p_k &= \frac{1}{2} \ln \left[\frac{\zeta^2}{s^{2m}} - 2 \frac{\zeta}{s^m} \cos 2km\pi + 1 \right]; \\ q_k &= \operatorname{arctg} \frac{\frac{\zeta}{s^m} - \cos 2km\pi}{\sin 2km\pi}; \quad p'_k = \frac{1}{2} \ln \left[\frac{1}{s^{2m}} - 2 \frac{1}{s^m} \cos 2km\pi + 1 \right]; \\ q'_k &= \operatorname{arctg} \frac{\frac{1}{s^m} - \cos 2km\pi}{\sin 2km\pi}; \quad \zeta = ms - m + 1. \end{aligned}$$

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a''_{II} — for the index $1/m$ of positive, odd

$$\begin{aligned}
 a_{II}^* = \ln \frac{x}{x-1} + x^{m-1} & \left[\ln \frac{1 - \frac{1}{x^m}}{1 - \frac{\zeta}{x^m}} + 2 \sum_{k=0}^{\frac{\frac{1}{m}-3}{2}} p_k \cos(2k+1)m\pi + \right. \\
 & + 2 \sum_{k=0}^{\frac{\frac{1}{m}-3}{2}} q_k \sin(2k+1)m\pi - 2 \sum_{k=0}^{\frac{\frac{1}{m}-3}{2}} p_k^* \cos(2k+1)m\pi - \\
 & \left. - 2 \sum_{k=0}^{\frac{\frac{1}{m}-3}{2}} q_k^* \sin(2k+1)m\pi \right], \quad (5-35)
 \end{aligned}$$

where

$$\begin{aligned}
 p_k &= \frac{1}{2} \ln \left[\frac{\zeta^2}{x^{2m}} + 2 \frac{\zeta}{x^m} \cos(2k+1)m\pi \right]; \\
 q_k &= \operatorname{arctg} \frac{\frac{\zeta}{x^m} + \cos(2k+1)m\pi}{\sin(2k+1)m\pi}; \quad q_k^* = \operatorname{arctg} \frac{\frac{1}{x^m} + \cos(2k+1)m\pi}{\sin(2k+1)m\pi}; \\
 p_k^* &= \frac{1}{2} \ln \left[\frac{1}{x^{2m}} + 2 \frac{1}{x^m} \cos(2k+1)m\pi + 1 \right].
 \end{aligned}$$

Exponential amplifier. For this amplifier integral (5-27) takes the form:

$$a_{II} = \int \frac{ds}{s(a)-1 - s \ln s} + C. \quad (5-36)$$

Final solution of integral (5-36)

$$\begin{aligned}
 a_{II} = \ln \frac{x}{x-1} + \frac{e^{-\frac{x-1}{a}}}{a} \ln \frac{-\frac{x-1}{a}}{\ln s - \frac{x-1}{a}} + \frac{e^{-\frac{x-1}{a}}}{a} & \left[\frac{-\frac{x-1}{a}}{1 \cdot 1!} + \frac{\left(-\frac{x-1}{a}\right)^2}{2 \cdot 2!} + \right. \\
 + \dots + \frac{\left(-\frac{x-1}{a}\right)^n}{n!} + \dots & \left. \right] - \frac{e^{-\frac{x-1}{a}}}{a} \left[\ln z - \frac{x-1}{a} + \frac{\left(\ln z - \frac{x-1}{a}\right)^2}{2 \cdot 2!} + \right. \\
 + \dots + \frac{\left(\ln z - \frac{x-1}{a}\right)^n}{n!} & \left. \right]. \quad (5-37)
 \end{aligned}$$

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Exponential amplifier. For it the integral (5-27) takes the form:

$$a_{11} = \int \frac{ds}{s(s) - \frac{\ln[(s-1)\zeta \ln M + 1]}{\zeta \ln M} + 1} + C. \quad (5-38)$$

Final solution of integral (5-38)

$$a_{11} = \ln \frac{s}{s-1} + e^p \left[-p + \frac{(-p)^2}{2 \cdot 2!} + \dots + \frac{(-p)^n}{n!} \right] + e^p \ln \frac{-p}{\ln(h-p)} - e^p \left[\ln(h-p) + \frac{[\ln(h-p)]^2}{2 \cdot 2!} + \dots + \frac{[\ln(h-p)]^n}{n!} \right], \quad (5-39)$$

where

$$p = (z+1)\zeta \ln M; \quad h = z\zeta \ln M - \zeta \ln M + 1 = (z-1)\zeta \ln M + 1.$$

Quasi-linear mode/conditions. Multistage amplifier actually can be performed with $b = \text{var}(\downarrow)$. Consequently, transient processes with the work of nonlinear cascade/stage in the quasi-linear mode/conditions make sense to examine only for the logarithmic and exponential amplifiers with $\beta < 1$. Taking into account the mating of the functional and quasi-linear sections of the characteristic, we obtain the final solutions of integral (5-27).

Logarithmic amplifier. For it

$$a_{III} = \ln \frac{x}{x-1} + \frac{a}{s} \ln \frac{d(x-1)}{x-d} + \frac{a}{d} \ln \frac{\frac{x}{d}-1}{\frac{x}{d} + \frac{a \ln d + 1 - a}{a} - \frac{x}{d}}. \quad (5-40)$$

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Exponential amplifier. For n of the positive even

$$\begin{aligned} a'_{III} = & \ln \frac{x}{x-1} + \frac{1}{hK_n} \ln \frac{h - nK_n^{\frac{1}{n-1}} + n-1}{h-s} + x^{\frac{1-n}{n}} \left[\ln \frac{1 + K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}}}{1 - K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}}} - \right. \\ & - \ln \frac{1 + x^{-\frac{1}{n}}}{1 - x^{-\frac{1}{n}}} - 2 \sum_{k=1}^{\frac{n}{2}-1} p_n \cos \frac{2k}{n} \pi + 2 \sum_{k=1}^{\frac{n}{2}-1} q_n \sin \frac{2k}{n} \pi + \\ & \left. + 2 \sum_{k=1}^{\frac{n}{2}-1} p'_n \cos \frac{2k}{n} \pi - 2 \sum_{k=1}^{\frac{n}{2}-1} q'_n \sin \frac{2k}{n} \pi \right], \quad (5-41) \end{aligned}$$

where

$$\begin{aligned} p_n &= \frac{1}{2} \ln \left[K_n^{\frac{1}{n-1}} x^{-\frac{2}{n}} - 2K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}} \cos \frac{2k}{n} \pi + 1 \right]; \\ q_n &= \operatorname{arctg} \frac{K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}} - \cos \frac{2k}{n} \pi}{\sin \frac{2k}{n} \pi}; \quad q'_n = \operatorname{arctg} \frac{x^{-\frac{1}{n}} - \cos \frac{2k}{n} \pi}{\sin \frac{2k}{n} \pi}; \\ p'_n &= \frac{1}{2} \ln \left[x^{-\frac{2}{n}} - 2x^{-\frac{1}{n}} \cos \frac{2k}{n} \pi + 1 \right]; \quad h = \frac{x}{K_n} + (n-1) \times \\ & \times (K_n^{\frac{1}{n-1}} - 1). \end{aligned}$$

For n of positive, odd

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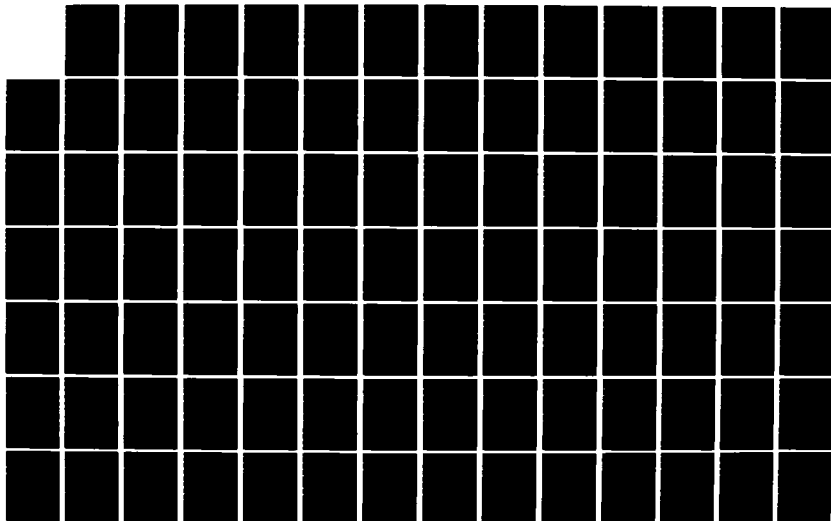
FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC
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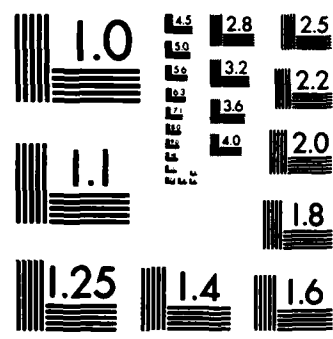
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$$\begin{aligned}
 a_{in}^* = & \ln \frac{x}{x-1} + \frac{1}{K K_n} \ln \frac{K - n K_n^{\frac{1}{n-1}} + n - 1}{K - s} x^{\frac{1-n}{n}} \left[\ln \frac{1 - x^{-\frac{1}{n}}}{1 - K^{\frac{1}{n-1}} x^{-\frac{1}{n}}} + \right. \\
 & + 2 \sum_{k=0}^{\frac{n-1}{2}} p_k \cos \frac{2k+1}{n} \pi - 2 \sum_{k=0}^{\frac{n-3}{2}} q_k \sin \frac{2k+1}{n} \pi - \\
 & \left. - 2 \sum_{k=0}^{\frac{n-3}{2}} p_k^* \cos \frac{2k+1}{n} \pi - 2 \sum_{k=0}^{\frac{n-1}{2}} q_k^* \sin \frac{2k+1}{n} \pi + 1 \right], \quad (5-42)
 \end{aligned}$$

where

$$\begin{aligned}
 p_n &= \frac{1}{2} \ln \left[K_n^{\frac{2}{n-1}} x^{-\frac{2}{n}} + 2 K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}} \cos \frac{2k+1}{n} \pi + 1 \right]; \\
 q_n &= \operatorname{arctg} \frac{K_n^{\frac{1}{n-1}} x^{-\frac{1}{n}} + \cos \frac{2k+1}{n} \pi}{\sin \frac{2k+1}{n} \pi}; \quad q_n^* = \operatorname{arctg} \frac{x^{-\frac{2}{n}} + \cos \frac{2k+1}{n} \pi}{\sin \frac{2k+1}{n} \pi}; \\
 p_n^* &= \frac{1}{2} \ln \left[x^{-\frac{2}{n}} + 2 x^{-\frac{1}{n}} \cos \frac{2k+1}{n} \pi + 1 \right].
 \end{aligned}$$

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Thus, expressions for the transient responses of amplifiers with LAKh (5-28), (5-29), (5-40) and SAKh with $\beta > 1$ and $\beta < 1$ (5-31), (5-32), (5-34), (5-41), (5-42) are obtained in the final form, and for the amplifier with EAKh and PAKh - in the form of converging series (5-37) and (5-38). It is possible to show that the law of the decrease of series/rows (5-37) and (5-38) when $\alpha \rightarrow \infty$ ($s \rightarrow s_1$) is proportional to ratio p/n , and at values $s = \gamma s_1$ - it is proportional to relation $\frac{p(1-\gamma)}{n}$, where $p = \frac{s-1}{s}$ - for the EAKh and $p = (x+1) \ln M$ -

for PAKh; z_y — value z , which corresponds to the steady-state mode/conditions with $\alpha \rightarrow \infty$.

In the real single-stage amplifiers with EAKh and PAKh it is possible to obtain maximum value of x_m of not more than 10. But even at values $x_m < 10$ for obtaining precise calculations it is necessary to take a sufficiently large number of members of series/row, on the order of 100 and more.

Transient responses $h = \frac{z}{z_y} = f(\alpha)$ of nonlinear cascade/stage with the work in the linear, logarithmic and quasi-linear modes/conditions, calculated by formulas (5-28), (5-29) and (5-40) with $a=1$, are depicted in Fig. 83a.

In calculation of the characteristics, we take the most probable range of LAKh of cascade/stage $d=10$. Transient responses for the quasi-linear mode of operation of cascade/stage are designed with $x=10$; 33 and 102, which corresponds to relative voltages on the input of the 1st, 2nd and 5th nonlinear cascades/stages at the end of the logarithmic range of five-stage amplifier.

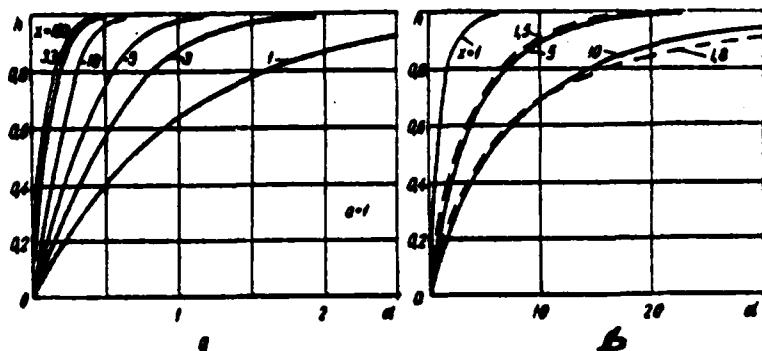


Fig. 83. The transient responses of the electron-tube nonlinear cascade/stage for the region of the short times: a) cascade/stage with LAKh; b) cascade/stage with MAC ($\beta > 1$); — $\beta = 2$; ---- $\beta = 5$.

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The transient responses of cascade/stage from MAC with $b = \text{var } (\uparrow)$, calculated by formulas (5-28), (5-34) and (5-35) for $\beta = 2$ and 5 with the work of cascade/stage in the linear and functional modes/conditions are shown in Fig. 83b. From the figures it is evident that with the increase of input time signal of establishment t_r and delay time t_d of impulse/momentum/pulse at the output of nonlinear cascade/stage with $b = \text{var } (\uparrow)$ are sharply reduced, and with $b = \text{var } (\uparrow)$ — they increase.

By time of establishment is understood the pulse rise-time from 0.1 to 0.9 of its maximal value.

In the functional amplifiers with $b = \text{var} (\downarrow)$, made on the tubes (inertia-free UP), with the shunting of load by semiconductor diodes of the type DG-Ts, D2 and D9 the impulses/momenta/pulses at the output of cascade/stage with the voltage at the input of more than 0.1 V have characteristic peak at the flat/plane apex/vertex. The value of peak depends on the value of input signal and grows with an increase in the latter. With the work of nonlinear cascade/stage in the functional mode/conditions the peak is not observed. It appears approximately/exemplarily in the middle of the quasi-linear section of the amplitude characteristic of the third nonlinear cascade/stage. In the three-stage amplifier with the logarithmic range 60 dB the peak appears in the middle of range and toward the end of it can achieve 30-40% of the value of impulse/momentum/pulse. The duration of peak does not exceed 0.3-0.4 μs .

During the amplification of impulses/momenta/pulses with a duration of $t_n > 0.5 \mu\text{s}$ the peak does not affect AKh of the amplifier. However, during the amplification of the impulses/momenta/pulses of the short duration, when $t_n < 0.5 \mu\text{s}$, the AKh of amplifier strongly differs from calculated logarithmic. The appearance of a peak in the impulse/momentum/pulse at the output of nonlinear cascade/stage depends on the inertness of semiconductor diode. If as the nonlinear

element/cell, which shunts anodic load of the UK, a germanium diode of the type DG-S1-DG-S4 or vacuum-tube diode is used, peak in the impulse/momentum/pulse at the output of nonlinear cascade/stage is absent with any input voltage. This means that germanium semiconductor diodes of the type DG-S1, DG-S2, DG-S3, DG-S4, DG-P3, DG-P4 and vacuum-tube diodes have smaller inertness in comparison with the diodes of the type DG-Ts1 - DG-Ts10.

Thus, for eliminating the peak it is necessary to apply the nonlinear elements/cells, which have the rapid response. Transient processes in the quite nonlinear element/cell, caused on finite time hole dislocations and electrons in the semiconductor, must last not more than hundredths of microsecond. These requirements satisfy germanium semiconductor diodes of the type DG-S1 - DG-S4 and DG-P3 - DG-P4.

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Inertial amplifier instrument.

Inertial UP includes the transistors. In the equivalent diagram (Fig. 82a) transistor on the common-base circuit

$$i_n(t) = \alpha_0(t) I_0 = \alpha_0 I_0 (1 - e^{-\frac{t}{\tau_0}}); \quad (5-43)$$

upon the start of transistor for common-emitter connection

$$i_n(t)_0 = \beta_0(t) I_0 = \beta_0 I_0 (1 - e^{-\frac{t}{\tau}}), \quad (5-44)$$

where I_0 and I_0 — jumps of input current of the cascade/stage, assembled on diagram in conformity with the OE and OB.

Between the nonlinear amplifier stages the emitter followers with low output resistance usually are connected. In connection with this transient processes in the nonlinear cascade/stage on the transistor can be investigated on the assumption that at the input of cascade/stage the voltage generator, which has internal resistor/resistance $R_r \ll R_{in}$ is connected. Without the large error it is possible to consider that the internal resistor/resistance of the source of signal is equal to zero, i.e., $R_r = 0$.

Then, passing from the input current generator to the input voltage generator and the slope/transconductance of the passage characteristic of transistor S_r [63], expression (5-43) and (5-44) it is possible to register in the general view

$$i_n(t) = S_r U_{in} (1 - e^{-\frac{t}{\tau}}), \quad (5-45)$$

where τ — time constant of transistor.

During the analysis of transient processes let us assume that the transistor is linearized with the help of the negative feedback.

In this case in equation (5-45) it is necessary to substitute the value $S_{z, \infty}$ and $\tau_{z, \infty}$ expressions for which are given in work [22].

Under the influence at the input of nonlinear transistor cascade/stage at the moment of time $t=0$ of voltage surge U_{sz} transient processes at zero time according to expressions (5-24) and (5-45) are described by the following nonlinear differential equation

$$\frac{dU_{sz}}{dt} + \frac{S_{z, \infty}}{C_0} U_{sz} = \frac{SU_{sz}}{C_0} (1 - e^{-\frac{t}{\tau}}). \quad (5-46)$$

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Linear conditions ($z \ll 1$). For the linear conditions equation (5-46) can be registered

$$\frac{dz}{dt} + bz = bz(1 - e^{-ct}), \quad (5-47)$$

where

$$b = \frac{1}{\tau} = \frac{1}{C_0 R_0}; \quad c = \frac{1}{\tau_r}.$$

For solving equation (5-47) it is possible to use the Laplace transform. Applying to equation (5-47) direct transformation of Laplace (5-3), we obtain

$$pz(p) + bz(p) = \frac{zcb}{p(p+c)},$$

whence

$$z(p) = \frac{zcb}{p(p+b)(p+c)}. \quad (5-48)$$

For finding the original $z(t)$ we apply inverse transformation of Laplace (5-4)

$$z(t) = \frac{1}{2\pi j} \int_{-j\infty}^{+j\infty} \frac{asbs^M}{p(p+b)(p+a)} dp. \quad (5-49)$$

Since pole images $z(p)$ are located in the left half-plane by complex variable p , for calculating the original $z(t)$ we apply the Cauchy theorem about the deductions. Lowering intermediate linings/calculations, finally we obtain

$$z(t) = x \left(1 - \frac{as^{-bt}}{s-b} + \frac{bs^{-at}}{s-a} \right). \quad (5-50)$$

With $t=0$ $z_0 = x$ and the transient response

$$h_{z,t} = \frac{z(t)}{z_0} = 1 - \frac{as^{-bt}}{s-b} + \frac{bs^{-at}}{s-a}. \quad (5-51)$$

Functional mode/conditions. Logarithmic amplifier with any foundation N . The differential equation, which describes transient processes,

$$\frac{dz}{dt} + me^{nz} = x(1 - e^{-q}), \quad (5-52)$$

where $n = \ln N = \frac{1}{\tau}$; $m = e^{-a} = e^{-\frac{1}{\tau}}$; $q = \frac{x}{\tau}$, can be solved approximately by one of the methods, enumerated in § 1.

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For finding the solution of equation (5-52) we introduce designation $p = e^{-nx}$. Then

$$\frac{dp}{dx} + pnx(1 - e^{-qx}) = nm. \quad (5-53)$$

We seek the solution in the form $p=rs$. As a result

$$s = Me^{-nx\left(1 + \frac{1}{q}e^{-qx}\right)},$$

$$r = nm \int e^{-1} dx + N = nm \int e^{-nx\left(1 + \frac{1}{q}e^{-qx}\right)} dx + N. \quad (5-54)$$

For simplification in the solution we accept $M=1$. For solving the integral we represent value s in the form of the series/row

$$s = e^{-nx} \left[1 - \gamma + \frac{\gamma^2}{2!} - \frac{\gamma^3}{3!} + \dots \right], \quad (5-55)$$

where $\gamma = \frac{nx}{q} e^{-qx}$.

In practice three cases can be met.

1. Inequality $q \gg 1$ is fulfilled. This corresponds to the case, when in the narrow-band and low-frequency amplifiers are used high-frequency transistors of the type P403--P416, connected on the diagram with OB, or transistors of the type P411, P418Zh, connected

on the diagram OB and OE. Then in expression (5-55) it suffices to take into account two first terms of series/row. Taking into account the done assumptions, and also the condition of the mating of the linear and logarithmic sections of AKh of cascades/stages during the determination of integration constant N into expressions (5-54), we obtain the following resultant expression for p:

$$p = rs = \left(1 - \frac{ns}{q}\right) \left[\frac{m}{s} (1 - e^{-ns(a-a_1)}) + \frac{n^2 ms}{q(ns-q)} e^{-a_1} (e^{-s(a-a_1)} - e^{-ns(a-a_1)}) + \left(1 - \frac{ns}{q} e^{-a_1}\right)^{-1} e^{-ns(a-a_1)+1} \right], \quad (5-56)$$

where $a_1 = \frac{1}{\tau}$ — value α , determined for this value of x from equation (5-50) at the moment/torque, when $z=1$.

In this case

$$s(a) = -\frac{1}{s} \ln p. \quad (5-57)$$

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Substituting expression (5-56) in (5-57), we have:

with $\alpha \rightarrow \alpha_1$,

$$z(\alpha) = 1;$$

with $\alpha \rightarrow \infty$

$$z(\alpha) = a \ln x + 1.$$

Transient response with the work of cascade/stage in logarithmic mode

$$h(\alpha) = \frac{z(\alpha)}{a \ln x + 1}. \quad (5-58)$$

2. Is performed exemplary/approximate equality $q \approx 1$, i.e., $\tau = \tau_p$. In this case series/row (5-55) - slowly converging. It descends slower, is the less α and the greater x . For the solution of problem it is necessary to consider a large number of terms, and integral (5-54) is not taken. For approximate solution of problem it is possible to use a Laplace transform or a complex conversion. Let us consider the solution of problem with the help of the Laplace

transform. For this equation (5-52) let us rewrite in the following form:

$$\frac{dz}{dt} + de^{nz} = xb(1 - e^{-ct}), \quad (5-59)$$

where $d = be^{-n}$.

On the basis of that presented in §1

$$W(p) = p; \quad H(p) = 1; \quad h(t) = L^{-1}H(p) = 1.$$

Then interval (5-9) for our case taking into account the mating of linear and logarithmic sections of AKh can be registered in the following form

$$z(t) = \int_0^t bx(1 - e^{-ct}) dt - \int_{t_1}^t de^{nz} dt, \quad (5-60)$$

which in accordance with expression (5-12) can be represented in the form of the recurrent series/rows

$$\begin{aligned} z_0(t) &= bx \left(\frac{ct^2}{2} - \frac{c^2 t^3}{3!} + \frac{c^3 t^4}{4!} - \dots \right); \\ z_1(t) &= z_0(t) - d(t - t_1) - dnbx \left[\frac{c(t^3 - t_1^3)}{3!} - \frac{c^2(t^4 - t_1^4)}{4!} + \frac{c^3(t^5 - t_1^5)}{5!} \right]; \\ z_2(t) &= z_0(t) - d(t - t_1) \left(1 + t_1 - \frac{t - t_1}{2!} \right) - dbx \left\{ \frac{c(t^3 - t_1^3)}{3!} - \right. \\ &\quad \left. - (dn + c) \left[\frac{c(t^4 - t_1^4)}{4!} - \frac{c^2(t^5 - t_1^5)}{5!} + \frac{c^3(t^6 - t_1^6)}{6!} \dots \right] + \right. \\ &\quad \left. + (t - t_1)(ct_1^3 - c^2 t_1^4 + c^3 t_1^5 - \dots) \right\}; \\ &\dots \dots \dots \end{aligned} \quad (5-61)$$

During the derivation of formulas (5-61) in expansion $e^{ns} = 1 + ns + \frac{(ns)^2}{2!} + \dots$ were considered only two members. This assumption is possible, since series/rows (5-61), beginning with $z_0(t)$, for the calculated intervals of time $t_1 < t < 3\tau$ they are rapidly converging. The calculation is obtained more precisely, the greater series/rows (5-61).

It is most expedient to calculate transient response with the use of series/rows (5-61) on a calculator.

3. Is fulfilled inequality $q \ll 1$, i.e., $\tau \gg t$. In this case transient processes are determined in essence by the inertial properties of transistor and have the temporary/time dependence, described by expression (5-45).

The transient responses of nonlinear cascade/stage with LAKh for two values of q with $x=1$ and 10, which corresponds to beginning and end/lead of LAKh in $d=10$, are depicted in Fig. 84. Experimental data coincide sufficiently well with the calculated ones.

Transient processes in the region of long times.

The distortion of flat/plane pulse apex is not that another as distortion of the transient response of the cascade/stage in the region of the long times, when inertness of UP it is possible to disregard and any UP it is possible to consider it inertia-free.

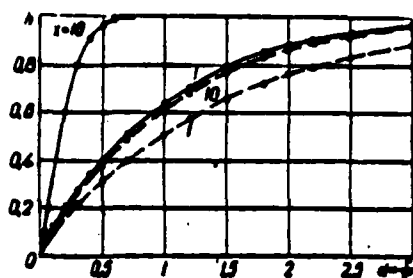


Fig. 84. The transient responses of transistor nonlinear cascade/stage with LAKh for the region of the short times: ----- $q=100$; --- $q=1$.

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The distortions of flat/plane pulse apex depend on the network elements, which are determining the frequency characteristic in the region of the lowest frequencies,. The equivalent diagram of the nonlinear cascade/stage for the region of the lowest frequencies (without taking into account the capacity/capacitance of filter) is depicted in Fig. 85.

For vacuum-tube amplifier $R'_{in} \approx R_o$, for the transistor with emitter follower — $R'_{in} = R_o = \frac{R_{o_1} R_{o_2}}{R_{o_1} + R_{o_2}}$.

For the most real values of the resistor/resistance to $R \geq 1$ kilohm and of the transient capacity/capacitance $C_1 \geq 0.05 \mu F$

(smallest capacitor C_1 for the vacuum-tube amplifiers) and the duration of pulse $t \leq 5-10 \mu s$, when inequality $RC_1 \gg t$ is fulfilled the current of the charge of capacities/capacitances C_1 and C_2 , which takes place through the nonlinear element/cell, during the action of impulse/momentum/pulse remains in effect constant. Consequently, resistor/resistance R_{non} of nonlinear element/cell for this value of the quantity of the voltage pulse is also constant, i.e., $R_{non} = r_1 = \text{const}$. Therefore the diagram, depicted in Fig. 85a, during the action of impulse/momentum/pulse with a sufficient degree of accuracy can be considered as linear. In this case one must take into account that to each value of input voltage corresponds its value of resistor/resistance R_{non} , determined from the appropriate formulas.

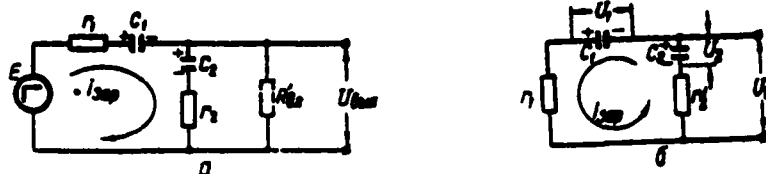


Fig. 85. The equivalent diagrams of the nonlinear cascade/stage for the region of the low frequencies: a) the process of the charge of capacities/capacitances; b) the process of the discharge of capacities/capacitances; $r_1 = \frac{R_{\text{max}} R}{R_{\text{max}} + R}$; $R = R_a$ and $R = R_n$ - respectively for the electron-tube and transistor amplifiers; $r_1 = R_{\text{max}}$; $R_{\text{bx}} = \frac{R_{\text{bx}} R'}{R_{\text{bx}} + R'}$; $R' = R_0$ and $R' = R_1$ - respectively for the electron-tube and the transistor amplifier; C and C_1 - with respect transient and separating capacities/capacitances.

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Distortion of flat/plane pulse apex.

The value, which most fully characterizes transient processes during the action of impulse/momentum/pulse, is relative decay in its flat/plane apex/vertex, the numerically equal to the ratio of absolute decay in the flat/plane apex/vertex ΔU toward the end of the action of impulse/momentum/pulse to maximum conservative value of the

impulse/momentum/pulse

$$\Delta = \frac{\Delta U}{U_{\text{MAX}}(0)} = \frac{U_{\text{MAX}}(0) - U_{\text{MAX}}(t_n)}{U_{\text{MAX}}(0)} = 1 - \frac{U_{\text{MAX}}(t_n)}{U_{\text{MAX}}(0)}, \quad (5-62)$$

where $U_{\text{MAX}}(0) = SU_{\text{IN}}R_0$ and $U_{\text{MAX}}(t_n)$ - output voltage/stress of impulse/momentum/pulse at the moments of time $t=0$ and $t=t_n$. Since the cascade/stage during the action of impulse/momentum/pulse is linear, for the analysis of transient processes in the diagram in Fig. 85a and determining the relative decay it is possible to use a Laplace transform.

Operational image of output potential of cascade/stage during the action of impulse/momentum/pulse according to diagram in Fig. 85a

$$U_{\text{MAX}}(p) = I_{\text{IN}}(p) Z_2(p) = \frac{E}{C_2(r_1 + r_2)(p + a)} \cdot \frac{1}{p + a} + \frac{E}{r_1 + r_2} \cdot \frac{p}{p + a}, \quad (5-63)$$

where $a = \frac{C_1 + C_2}{C_1 C_2 (r_1 + r_2)}$; $E = SU_{\text{IN}}R_0$.

Original of the voltage/stress

$$U_{\text{MAX}}(t) = E \left[\frac{C_1}{C_1 + C_2} - \left(\frac{C_1}{C_1 + C_2} - \frac{r_1}{r_1 + r_2} \right) e^{-at} \right]. \quad (5-64)$$

Substituting in expression (5-62) of value $U_{\text{MAX}}(0)$ and $U_{\text{MAX}}(t_n)$, we obtain expression for the relative decay, caused by the charge of capacities/capacitances C_1 and C_2 ,

$$\Delta = 1 - \frac{C_1(r_1 + r_2)}{r_1(C_1 + C_2)} + \left[\frac{C_1(r_1 + r_2)}{r_1(C_1 + C_2)} - 1 \right] e^{-at_n}. \quad (5-65)$$

With fulfilling of inequality $\alpha t_n \ll 1$ expression (5-65) is simplified

$$\Delta = t_n \left[\frac{C_1 + C_2}{C_1 C_2 (r_1 + r_2)} - \frac{1}{C_2 r_2} \right]. \quad (5-66)$$

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Substituting in expression (5-66) the value of resistor/resistance of r_2 ,

$$r_2 = R_{\text{max}} = \frac{R_0}{\varphi(s) - 1} = \frac{R_0}{\frac{s}{s} - 1},$$

we obtain

$$\Delta_1 = t_n [\varphi(s) - 1] \left[\frac{C_1 + C_2}{C_1 C_2 \varphi(s) (r_1 + R_0)} + \frac{1}{C_2 R_0} \right]. \quad (5-67)$$

The resulting (total) relative decay in the flat/plane pulse apex at the output of the nonlinear cascade/stage

$$\Delta = \Delta_1 + \Delta_2 + \Delta_3 - \Delta_4, \quad (5-68)$$

where Δ_1 - relative decay, capacities/capacitances C_n forming with the charge in the cathode circuit of tube (or C_e in the emitter circuit of transistor); Δ_2 - relative decay, which is formed with the

charge of capacity/capacitance in the circuit of the screen grid of tube; Δ_1 - relative decay, which is formed with the charge of the capacity/capacitance of the filter, connected in series with the load.

Values Δ_1 , Δ_2 , and Δ_3 over the dynamic range are constant and are determined from formulas for linear amplifier [56].

With the increase of input voltage component Δ_1 rapidly grow also many times exceed remaining components/terms/addends of sum (5-68). Therefore during the calculation of relative decay at the output of n-cascade functional amplifier for the nonlinear cascade/stage, assembled on the diagram (Fig. 82), without the considerable error it is possible to take

$$\Delta \approx \Delta_1.$$

After the break-down of impulse/momentum/pulse capacity/capacitance C , is discharged through two parallel-connected resistors/resistances of R and R_{out} . Transient processes during the discharge of capacity/capacitance C , are described by nonlinear differential equation (5-24). In view of the fact that the form of decay in the impulse/momentum/pulse does not have vital importance, the detailed analysis of transient processes during the discharge of capacity/capacitance C , is not given. It should be noted that as a

result of an increase of resistor/resistance R_{res} in the process of the discharge of capacity/capacitance C_0 decay in the impulse/momentum/pulse is somewhat stretched in comparison with the pulse. As shown by the theoretical and experimental front of the research, carried out by the author, the decay time in the impulse/momentum/pulse at the output of cascade/stage with $b=\text{var}$ (\downarrow) with $x \geq d$ is 3-5 times more than the time of establishment.

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Parasitic reverse/inverse selection.

The value, which most fully characterizes transient processes in the nonlinear cascade/stage after the break-down of impulse/momentum/pulse, is the relative overshoot, numerically equal to the ratio of the voltage/stress of overshoot U_o to the maximum value of the impulse/momentum/pulse

$$\beta = \frac{U_o}{U_{\text{max}}(0)}. \quad (5-09)$$

General/common/total relative overshoot is composed of the same various components

$$\beta = \beta_1 + \beta_2 + \beta_3 - \beta_4$$

caused by discharge capacities/capacitances, as for Δ_1 , Δ_2 , Δ_3 and Δ_4 . Components β_1 , β_2 and β_3 are the same, as for the linear

cascade/stage, they are determined from the same formulas and they are constant over the dynamic range. The value of the overshoot β_1 , which is formed due to the discharge of the transient capacitor C_1 , is changed over the dynamic range and can reach the significant magnitudes.

For determining the voltage/stress U_s , we will use the equivalent diagram of the discharge of capacities/capacitances C_1 and C_2 , depicted in Fig. 85, where through r' , the resistor/resistance of nonlinear element/cell for voltage/stress U_s is marked. If as the nonlinear element/cell are used two diodes, connected by different polarities (Fig. 30e), maximum voltage/stress U_s , actually composes tens of millivolts and resistor/resistance r' , reaches the units of kilohm.

During the determination of voltage/stress U_s at zero time of the discharge of capacities/capacitances the diagram (Fig. 85b) can be considered linear. For simplification in the analysis let us take $r'_1 < R'_m$. Then

$$U_s = \frac{U_1 r'_1}{r_1 + r'_1} + \frac{U_2 r'_2}{r_2 + r'_2} - U_s = U_1 \frac{r'_1}{r_1 + r'_1} - U_s \left(1 - \frac{r'_1}{r_1 + r'_1}\right), \quad (5-70)$$

where $U_1 = \frac{EC_1}{C_1 + C_2} (1 - e^{-\omega t})$; $U_2 = \frac{EC_2}{C_1 + C_2} (1 - e^{-\omega t})$ - respectively

voltage/stress, to which capacities/capacitances C_1 and C_2 toward the end of the action of impulse/momentum/pulse will be loaded by

duration t_p . Substituting their values into expression (5-70), we obtain

$$U_0 = E(1 - e^{-at_p}) \left[\frac{C_1}{C_1 + C_2} \cdot \frac{r'_2}{r_1 + r'_2} - \frac{C_1}{C_1 + C_2} \left(1 - \frac{r'_2}{r_1 + r'_2} \right) \right].$$

In this case the relative overshoot

$$\beta_1 = \frac{r_1 + r_2}{r_2(C_1 + C_2)} \left[C_2 \frac{r'_2}{r_1 + r'_2} - C_1 \left(1 - \frac{r'_2}{r_1 + r'_2} \right) \right] (1 - e^{-at_p}). \quad (5-71)$$

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With fulfilling of inequality $at_p \ll 1$

$$\beta_1 = \frac{t_p}{C_1 C_2 r_2} \left[C_2 \frac{r'_2}{r_1 + r'_2} - C_1 \left(1 - \frac{r'_2}{r_1 + r'_2} \right) \right]. \quad (5-72)$$

The curves of change $\Delta_1 = f(x)$ and $\beta_1 = f(x)$ in the dynamic range for the logarithmic cascade/stage at the different values of capacities/capacitances C_1 and C_2 are given in Fig. 86. During the calculation it is accepted: $R_0 = 2$ kilohm; $d = 10$. From the figure one can see that, varying with the values of capacities/capacitances C_1 and C_2 , it is possible to considerably decrease values Δ_1 and β_1 over entire dynamic range. Unbroken curve 1 corresponds to the case of recorrection of flat/plane pulse apex. Curves 3, 4 are optimum.

MULTISTAGE AMPLIFIER.

Distortion of pulse edge.

The differential equation, which describes transient processes at the output of the i nonlinear cascade/stage of n -cascade video amplifier at the initial moment of acting the impulse/momentum/pulse,

$$\frac{dz_i}{d\alpha_0} + \varphi(z_i) z_i = K_n z_{i-1}(\alpha_0), \quad (5-73)$$

where z_{i-1} - relative output potential of $(i-1)$ nonlinear cascade/stage.

The transient responses of logarithmic amplifier ($b = \text{var}(\downarrow)$) for different number of nonlinear cascades/stages with $N=2.72$ and $d=10$ are depicted in Fig. 87.

Characteristics are designed graph-analytically according to equation (5-73) with any n for the end/lead of the logarithmic range of n -cascade amplifier. This corresponds to relative voltage on the input of the first cascade/stage $x=d$.

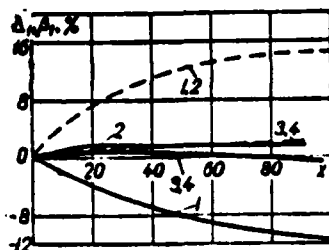


Fig. 86. The curves of change Δ_1 and β_1 in the dynamic range for the logarithmic cascade/stage: ---- $\Delta_1=f(x)$; ---- $\beta_1=\phi(X)$; $1-c_1=c_1=0.1 \mu F$; 2 - $C_1=0.1 \mu F$; $C_2=5 \mu F$; 3 - $C_1=1 \mu F$; $C_2=5 \mu F$; 4 - $c_1=c_2=5 \mu F$.

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Being congruent/equating the transient responses, given in Fig. 87, with the characteristics of n-cascade linear amplifier, given in work [29], let us compose Tables 11 of relative change $t_{y, \text{nor}}$ and $t_{s, \text{nor}}$ amplifier with the input voltage, which corresponds to the end/lead of the logarithmic range in comparison with $t_{y, \text{min}}$ and $t_{s, \text{min}}$ when amplifier reinforces low signals and operates in the linear conditions. From Table 11 it is evident that with $n \geq 5$ value $\zeta = \frac{t_{y, \text{nor}}}{t_{y, \text{min}}}$ and $\delta = \frac{t_{s, \text{nor}}}{t_{s, \text{min}}}$ they remain in effect constant.

Analogously it is possible to calculate transient responses, also, for other types of multistage FU with $b=\text{var}(\downarrow)$.

ff
Relative decay in the pulse apex and relative overshoot.

Relative decay in the flat/plane pulse apex at the output of
n-cascade functional video amplifier

$$\Delta_0 = \sum_{i=1}^n \Delta_{i \text{ MAX}}, \quad (5-74)$$

where $\Delta_{i \text{ MAX}}$ - component of general/common/total relative decay at the output of video amplifier, caused on the decay, which are formed in the i nonlinear cascade/stage. Formula (5-74) is accurate with fulfilling of inequality $\Delta_{i \text{ MAX}} < 10-15\%$.

Relative decay $\Delta_{i \text{ MAX}}$ has different expressions with the work of nonlinear cascades/stages in different modes/conditions.

Table 11.

Таблица 11

n	1	2	3	4	5
ζ	0,077	0,119	0,097	0,084	0,0835
δ	0,24	0,154	0,117	0,098	0,09

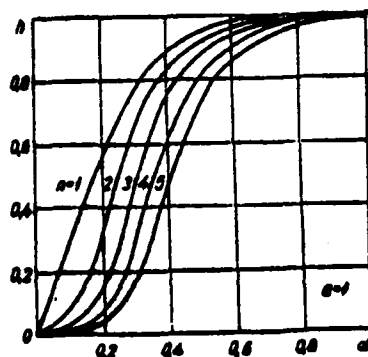


Fig. 87. Transient responses of multistage logarithmic amplifier.

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With the work of cascades/stages in functional and quasi-linear modes/conditions $\Delta_{i, \text{max}}$ it is determined by expression (5-67), since value Δ_{ii} of the i cascade/stage is transmitted to the output of amplifier by the subsequent cascades/stages, which work in the quasi-linear mode/conditions, without the change.

With the work of the i cascade/stage in the linear conditions the general/common/total expression for $\Delta_{11 \text{ BMAX}}$, the transmitted to the output functional amplifier, can be registered in the following form:

$$\Delta_{11 \text{ BMAX}} = \frac{Z(U_{\text{BMAX}}) - Z(U_{\text{BMAX}}(1 - \Delta_1))}{Z(U_{\text{BMAX}})}. \quad (5-75)$$

Using expressions for FAKh, given in Table 1, on the basis of expression (5-75) we can register:

for the logarithmic amplifier

$$\Delta_{11 \text{ BMAX}} = -\frac{\ln(1 - \Delta_1)}{\ln X + 1}, \quad (5-76)$$

or with $\Delta_1 \leq 5\%$

$$\Delta_{11 \text{ BMAX}} = \frac{\Delta_1}{\ln X + 1} = \frac{\Delta_1}{2},$$

where X and Z - standardized/normalized voltages/stresses of amplifier, which correspond to the moment/torque of the work of the i cascade/stage in the linear conditions;

for the amplifier with SAKh with $\beta = 1/n < 1$

$$\Delta_{11 \text{ BMAX}} = \frac{n \sqrt[n]{X} (1 - \sqrt[n]{1 - \Delta_1})}{2}. \quad (5-77)$$

The greatest decay Δ_1 is observed at the end of the dynamic

range of FAKh and is determined by expression (5-74), in which the components are calculated from formula (5-67). As can be seen from Fig. 86 at the unsuccessfully selected values of capacitors C_1 and C_2 , general/common/total decay can achieve the significant magnitude. Curves 3, 4 are the most optimum case of the selection of capacities/capacitances for the multistage amplifier, made on the diagram in Fig. 82, when in the part of the cascades/stages is observed overcorrection of flat/plane pulse apex, and in other - undercorrection. As a result at the output of amplifier the mutual compensation for the distortion of flat/plane pulse apex occurs.

Under the influence on the input n-cascade FU of ideal impulse/momentum/pulse without the reverse/inverse overshoot relative overshoot at the output of the amplifier

$$\beta_o = \sum_{i=1}^n \beta_{i \max}, \quad (5-78)$$

where $\beta_{i \max}$ - component of general/common/total relative overshoot at the output of amplifier, caused on the overshoot, which are formed in the i nonlinear cascade/stage.

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Formula (5-78) is valid [56] with fulfilling of inequality $\beta_{i \max} < 10-15\%$.

Functional amplifiers with $b = \text{var}(\downarrow)$ possess the property of accentuation of reverse/inverse overshoots. This is caused by the fact that the parasitic overshoots, which are formed in the nonlinear cascades/stages, many times of less than the signal are reinforced according to the linear law with $b = \text{makc} = \text{const}$, and signals are reinforced according to the functional law with $b = \text{var}(\downarrow)$. As a result, at the output of FU toward the end of the dynamic range 60-80 dB relative overshoot can reach to 30-60% [7]. Negative property of FU with $b = \text{var}(\downarrow)$ - to stress reverse/inverse overshoots - is exhibited also when video amplifier itself is ideal and in it reverse/inverse overshoots (ideal video amplifier) are not formed, but its input real impulses/momenta/pulses with the insignificant reverse/inverse overshoot enter. Thus, for instance, with the relative overshoot at the input of ideal logarithmic video amplifier $\beta_{\text{in}} = 0.01\%$ relative overshoot at the output of video amplifier toward the end of the dynamic range 80 dB (with $a=1$) reaches $\beta_{\text{out}} = 10\%$; with $\beta_{\text{in}} = 0.1\%$ $\beta_{\text{out}} = 32\%$ and with $\beta_{\text{in}} = 1\%$ $\beta_{\text{out}} = 54\%$.

With the decrease of coefficient of a of value β_{out} they grow.

On the basis of the analysis conducted it is possible to do a following conclusion. For decreasing the parasitic reverse/inverse

overshoot at the output of functional video amplifier it is necessary to in every possible way reduce the overshoots, which are formed both in the nonlinear cascades/stages of video amplifier and in the amplifier circuit, connected before the video amplifier.

Considerable parasitic reverse/inverse overshoot is the fundamental reason, which limits the wide application of functional video amplifiers. The elimination of parasitic overshoots is the fundamental problem, with solution of which the field of application of functional video amplifiers with $b = \text{var}(\downarrow)$, considerably will be widened.

WAYS OF DECREASING THE DISTORTIONS OF PULSE SIGNAL IN FUNCTIONAL VIDEO AMPLIFIERS.

As a result of the conducted investigations by the author some circuit solutions are determined, with the help of which it is possible to decrease the decay and flat/plane pulse apex and parasitic overshoot at the output of nonlinear cascade/stage, caused on transient capacity/capacitance.

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To such solutions should be related: the correction of the

help of the separating capacity/capacitance C_s ; insertion of nonlinear element ahead of depletion-layer capacitance C_d ; load shunting by type 1 and type 2 nonlinear dividers; inclusion/connection of the nonlinear element/cell, which forms FAKh, into the circuit of consecutive negative current feedback (cathode or emitter).

With the shunting of load by nonlinear dividers simultaneously occurs the stabilization of the delay time of signal over the dynamic range.

All enumerated solutions are in sufficient detail examined in work [7]. However, the case of including the nonlinear element/cell to the transient capacity/capacitance work indicated examines without taking into account the effect of the separating capacity/capacitance C_s , what cannot be made for the transistor amplifiers.

Inclusion/connection of nonlinear element/cell to the transient capacity/capacitance.

The equivalent diagrams of the cascade/stage for the region of the lowest frequencies are depicted in Fig. 88.

During the action of impulse/momentum/pulse we consider diagram linear, since for the real cases for practice is fulfilled inequality $C_s r_1 \gg t_m$.

Distortion of flat/plane pulse apex. Image of output voltage/stress according to diagram (Fig. 88a)

$$U_{\max}(p) = EBC_1 r_1 \left[\frac{p}{p^2 + Mp + B} + m \frac{p^2}{p^2 + Mp + B} \right], \quad (5-79)$$

where

$$B = \frac{1}{C_1 C_2 (r_1 r_2 + r_1 r_3 + r_2 r_3)}; \quad M = \frac{C_1 r_1 + C_1 r_2 + C_2 r_3}{C_1 C_2 (r_1 r_2 + r_1 r_3 + r_2 r_3)}; \quad m = C_2 r_3.$$

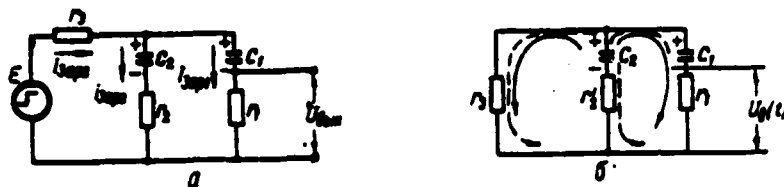


Fig. 88. Equivalent diagrams of the cascade/stage for the region of the lowest frequencies with connection of nonlinear element/cell to the transient capacitor C_1 : a) the diagram of the charge of capacities/capacitances; b) the diagram of the discharge of capacities/capacitances; $r_1 = R_{11}$; $r_2 = R_{22}$;

$$\frac{1}{r_1} = \frac{1}{R} + \frac{1}{R_{max}} \approx \frac{1}{R}; r_2 = R_{max}.$$

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Original of the output voltage/stress

$$U_{max}(t) = EBC_1 r_1 \left[\frac{1}{b} e^{-\frac{Mt}{2}} \text{sh } bt + me^{-\frac{Mt}{2}} \left(\text{ch } bt - \frac{M}{2b} \text{sh } bt \right) \right], \quad (5-80)$$

where

$$b = \sqrt{\frac{M^2}{4} - B}.$$

Substituting in expression (5-62) value $U_{max}(t)$ with $t=0$ and we obtain expression for relative decay in the flat/plane apex/vertex

$$\Delta = 1 - \left[\frac{1}{b m} \text{sh } b t_n + \text{ch } b t_n - \frac{M}{2b} \text{sh } b t_n \right] e^{-\frac{M t_n}{2}}. \quad (5-81)$$

Voltage/stress of parasitic reverse/inverse overshoot at the initial moment of the discharge of capacities/capacitances C_1 and C_2 according to diagram in Fig. 88b

$$U_2 = \frac{U_1}{1 + \frac{r_2' r_2}{r_1(r_2' + r_2)}} - \frac{U_2}{1 + \frac{r_2'(r_1 + r_2)}{r_1 r_2}}. \quad (5-82)$$

Image of voltage/stress U_1 according to Fig. 88a

$$U_1(p) = I_{\text{exp}}(p) Z_1(p) = EB \left[\frac{1}{p^2 + Mp + B} + m \frac{p}{p^2 + Mp + B} \right].$$

Original of the voltage/stress

$$U_1(t) = E \left[1 - e^{-\frac{Mt}{2}} \left(\text{ch } bt + \frac{M}{2b} \text{sh } bt + \frac{mB}{b} \text{sh } bt \right) \right]. \quad (5-83)$$

It is analogous, we obtain for voltage/stress U_2 on capacity/capacitance C_2

$$U_2(t) = E \left[1 - e^{-\frac{Mt}{2}} \left(\text{ch } bt + \frac{M}{2b} \text{sh } bt + \frac{nB}{b} \text{sh } bt \right) \right], \quad (5-84)$$

where $n = C_1 r_1$.

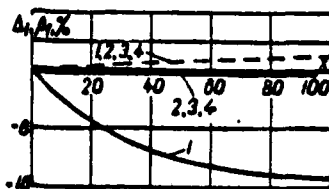


Fig. 89. The curves of change of Δ_1 and β_1 in the dynamic range of the logarithmic cascade/stage, made on the diagram in Fig. 88.

Capacitance values are the same as in Fig. 86.

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Substituting expression (5-82) in (5-68), taking into account conditions (5-83) and (5-84) when $t = t_n$, we obtain

$$\beta_1 = A \left[\frac{1 - e^{-\frac{Mt_n}{2}} \left(\text{ch } bt_n + \frac{M}{2b} \text{sh } bt_n + \frac{nB}{b} \text{sh } bt_n \right)}{1 + \frac{r_2' r_3}{r_1 (r_2' + r_3)}} - \frac{1 - e^{-\frac{Mt_n}{2}} \left(\text{ch } bt_n + \frac{M}{2b} \text{sh } bt_n + \frac{nB}{b} \text{sh } bt_n \right)}{1 + \frac{r_2' (r_1 + r_3)}{r_1 r_3}} \right], \quad (5-85)$$

where $A = \frac{r_1 r_3 + r_1 r_2' + r_2' r_3}{r_1 r_3}$.

Dependences $\Delta_1 = f(x)$ and $\beta_1 = \phi(x)$ for the logarithmic cascade/stage upon the inclusion/connection of nonlinear element/cell to the transient capacity/capacitance are given in Fig. 89. From these curves it is evident that during the appropriate selection of

capacities/capacitances C_1 and C_2 , it is possible to obtain the virtually constant values Δ_1 and β_1 , with a change of the signal in entire dynamic range.

§3. Transient in the selective amplifiers with the nonlinear load.

Transient processes in the tuned amplifier.

During the research of transient processes in selective amplifiers let us assume that UP is linear in entire dynamic range. This assumption in practice usually is performed. As a result of both one cascade/stage and multistage FU upon the inclusion at the moment of time $t=0$ of voltage surge

$$u_{\text{BX}}(t) = U_{\text{BX}}(t) \cos \omega t,$$

where $U_{\text{BX}}(t)$ - envelope of input voltage.

Then output potential

$$u_{\text{BHX}}(t) = U_{\text{BHX}}(t) \sin [\omega t - \psi(t)],$$

where $U_{\text{BHX}}(t)$ - envelope of output voltage/stress, which is the slowly varying function of time; $\psi(t)$ - slowly varying phase.

Since during the research of transient processes is of interest only envelope of oscillations, i.e., the law of a change in the amplitude of oscillations at the output of amplifier, as shown in work [33], it is most expedient to investigate transient processes by obtaining the approximate differential equations for the envelopes (shortened equations) with their subsequent solution by graphoanalytical method. With this method of study the problem is simplified, since interesting us envelopes are obtained without the plotting of curves of high frequency, and, furthermore, abbreviated equations for the envelope are the nonlinear differential equations of lower order than the initial differential equations of transient process.

Furthermore, for the simplest diagrams differential equations can be solved in the final form.

The detailed methodology of the composition of the shortened equations is given in work [29]. In accordance with this methodology let us compose the shortened equation of cascade/stage with the single duct/contour, shunted by the nonlinear element/cell (Fig. 32),. Complex transmission factor of the cascade/stage

$$K(j\omega) = SZ(j\omega) = \frac{j\omega LS}{(j\omega)^2 LC_0 + j\omega L\epsilon_{0, \text{max}} + 1}. \quad (5-86)$$

Taking into account that equivalent circuit damping

$$\beta_0 = \frac{R_{e. max}}{\omega_0 L_0} = \omega_0 L_0 g_{e. max} \quad (5-87)$$

we obtain

$$K(j\omega) = \frac{S}{g_{e. max}} \cdot \frac{1 - \frac{\omega}{\omega_0} \beta_0}{\left(\frac{j\omega}{\omega_0}\right)^2 + \frac{j\omega}{\omega_0} \beta_0 + 1}.$$

Let us find expression for the shortened complex amplification factor with a small detuning, when $\omega = \omega_0 + \Delta\omega$. Taking into account that

$$\frac{\omega}{\omega_0} = 1 + \frac{\Delta\omega}{\omega_0} = 1 + \frac{\delta}{\omega_0},$$

we obtain

$$K[j(\omega_0 + \Delta\omega)] = \frac{S}{g_{e. max}} \cdot \frac{\beta_0 + j \frac{\Delta\omega}{\omega_0} \beta_0}{\left(\frac{j\Delta\omega}{\omega_0}\right)^2 + 2j \frac{j\Delta\omega}{\omega_0} + \beta_0 + j \frac{\Delta\omega}{\omega_0} \beta_0}.$$

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Relative detuning $\Delta\omega/\omega_0$ and attenuation δ are the small first-order quantities. Rejecting in the numerator and the denominator all terms of the second and higher than the order of smallness and substituting the value β_0 , we obtain

$$K(j\Delta\omega) = \frac{S}{2j\Delta\omega C_0 + g_{e. max}} \quad (5-88)$$

There is in effect greatest interest in the case, when tuned amplifier is tuned to a frequency of signal. Taking into account this, it is possible to use not with complex, but real envelopes. Then enveloping at the output

$$U_{\text{BHX}}(t) = K(j\Delta\omega) U_{\text{BX}}(t).$$

Substituting the value of $K(j\Delta\omega)$, we obtain

$$(2j\Delta\omega C_0 + g_{\text{B. HON}}) U_{\text{BHX}}(t) = S U_{\text{BX}}(t).$$

Considering $j\Delta\omega$ as the differential operator $j\Delta\omega = d/dt$, we obtain the shortened equation for the envelope

$$2C_0 \frac{dU_{\text{BHX}}}{dt} + g_{\text{B. HON}} U_{\text{BHX}} = S U_{\text{BX}}^* \quad (5-89)$$

FOOTNOTE ¹. In equation (5-89) by U should be understood the amplitude values of voltages/stresses. ENDFOOTNOTE.

For the low-frequency gauge (transistor) in equation (5-89) it is necessary to substitute the value of slope/transconductance at the resonance frequency of oscillatory circuit.

For obtaining by required AKh of cascade/stage the conductivity for the fundamental harmonic of anode (collector) current in equation

(5-89) is described by expression (5-25), in which

$$g_0 = g_{\text{max}} + g_{0.0} + g_m + g_{\text{sx}} = \frac{1}{R_0},$$

where g_m - conductivity of shunt.

Substituting into equation (5-89) conductivity g_0 , relative time α and the standardized/normalized voltages/stresses x and z , we obtain

$$\frac{dz}{d\alpha} + \varphi(z)z = x(\alpha), \quad (5-90)$$

where

$$\alpha = \frac{t}{2C_0R_0}. \quad (5-91)$$

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Equation (5-90) is correct for the amplifiers with the inertia-free and with inertial UP, since standardized values x and z do not depend on the absolute value of the slope/transconductance of passage characteristic of UP.

Equation (5-90) is analogous (with 5-26), that describes transient processes in aperiodic FU. Therefore, all solutions, obtained in §2 during the analysis of transient processes in aperiodic FU, during the replacement of value α on α are valid also

for resonance FU.

Comparing the relative time of video amplifier (aperiodic amplifier) α , with the relative time of tuned amplifier with the single ducts/contours α , we see that with the fulfillment of equality $\alpha = \alpha$ the value of resistors/resistances R_0 in the anode circuits of UK of video amplifier is 2 times more than in the cascades/stages of tuned amplifier (with the identical capacities/capacitances in the anode circuits C_0). Consequently, in this case logarithmic video amplifier with the work in the linear conditions has a passband 2 times less, and maximum amplification factor 2 times larger than tuned amplifier. The distortions of pulse edge in both amplifiers are identical. With the equality of ranges of LAKh absolute changes in the delay time of signal t , and time of the establishment of impulse/momentum/pulse t , in this case in both amplifiers are identical.

With the equality of stray capacitances C_0 and resistors/resistances R_0 in the anode circuits of the cascades/stages of video amplifier and tuned amplifier the maximum factors of amplification and passband of both amplifiers are identical. Then with the equality of ranges of LAKh relative changes in the time lag of signal $\Delta\alpha$ in the aperiodic and tuned amplifiers identical, but absolute a change of the delay time of signal in logarithmic video

amplifier $\Delta t_{e, \dots}$ are 2 times less than in the resonance, since

$$\Delta t_{e, \dots} = \Delta \alpha_e R_o C_o; \quad (5-92)$$

$$\Delta t_{e, \dots} = \Delta \alpha 2 R_o C_o, \quad (5-93)$$

then when $\Delta \alpha_e = \Delta \alpha$

This property of logarithmic amplifiers should be considered during the design of different equipment.

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TRANSIENT PROCESSES IN BANDPASS AMPLIFIER.

For finding the equation, which describes transient processes, we will use the expression of the abbreviated transmission factor of band-pass amplifier (2-157), which for the optimum parameters

$$K(j\Omega) = K_n \frac{-j\delta_1^2 (q_n + A^2)}{4 \left(\frac{j\Omega}{\omega_0} \right)^2 + 2(\delta_1 + \delta_{2n}) \frac{j\Omega}{\omega_0} + [\delta_1 \delta_{2n} + 0.5(\delta_1^2 + \delta_{2n}^2)]}, \quad (5-94)$$

where

$$\delta_{2n} = \delta_1 \varphi(z); \quad (5-95)$$

$$\varphi(z) = (q_n + A^2) \frac{z}{i} - A^2. \quad (5-96)$$

We write/record the equation, which connects the output and

input voltage through the shortened symbolic transmission factor

$$U_{\text{MAX}} = K(j\Omega) \dot{U}_{\text{BX}}, \quad (5-97)$$

or

$$\left\{ 4 \left(\frac{j\Omega}{\omega_0} \right)^2 + 2(\delta_1 + \delta_{20}) \frac{j\Omega}{\omega_0} + [\delta_1 \delta_{20} + 0,5(\delta_1^2 + \delta_{20}^2)] \right\} U_{\text{MAX}} = \\ = K_u \delta_1^2 (q_u + A^2) U_{\text{BX}}. \quad (5-98)$$

Factor j , which indicates phase displacement between the output and input voltage on 180° , in the right side of equation (5-94) is omitted, since only envelopes interest us. Substituting in equation (5-98) $(j\Omega)^2$ on d^2/dt^2 , $j\Omega$ on d/dt , we obtain the shortened differential equation, which describes transient processes in the nonlinear cascade/stage

$$4 \frac{d^2 U_{\text{MAX}}}{dt^2} + 2\omega_0 (\delta_1 + \delta_{20}) \frac{dU_{\text{MAX}}}{dt} + \omega_0^2 [\delta_1 \delta_{20} + 0,5(\delta_1^2 + \delta_{20}^2)] U_{\text{MAX}} = \\ = K_u \delta_1^2 \omega_0^2 (q_u + A^2) U_{\text{BX}}. \quad (5-99)$$

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Passing to the standardized/normalized voltages/stresses x , z , relative time α and taking into account expression (5-95), (5-96), finally we obtain

$$\frac{d^2 x}{d\alpha^2} + [1 + \varphi(z)] \frac{dx}{d\alpha} + (\varphi(z) + 0,5[1 + \varphi^2(z)]) x = (q_u + A^2) x, \quad (5-100)$$

where

$$\alpha = \frac{\omega_0}{2} t. \quad (5-101)$$

The transient responses of nonlinear cascade/stage of FU for different levels of the input standardized/normalized voltage can be designed according to equation (5-100) by graphic method or in calculator.

The transient responses of nonlinear cascade/stage with the optimum the parameters, that works in the linear, logarithmic and quasi-linear mode/conditions, are given in Fig. 90. Characteristics are constructed graphically according to A. D. Bashkirov's method. During the calculation it is accepted: $d_1=10$; $q_1=0.6$; $\phi=0.7$. This corresponds to critical coupling between the ducts/contours in the work of cascade/stage in the linear conditions.

Being congruent/equating the transient responses, given in Fig. 83 and 90, we see that in the band-pass amplifier with the optimum parameters the delay time over the dynamic range is more stable than in the tuned amplifier with the single ducts/contours. It is obvious that in the multistage functional band-pass amplifier with the optimum parameters the delay time of signal over the dynamic range will be also stabilized.

The transient response of the i nonlinear cascade/stage of n -cascade amplifier can be designed graphically (or it is designed in calculator) according to the equation

$$\frac{d^2 z_i}{ds^2} + [1 + \varphi(z_i)] \frac{dz_i}{ds} + \{\varphi(z_i) + 0,5[1 + \varphi^2(z_i)]\} z_i = (q_n + A^2) K_n z_{i-1} \quad (5-102)$$

by the consecutive construction of the transient responses of each of the cascades/stages, beginning from the first.

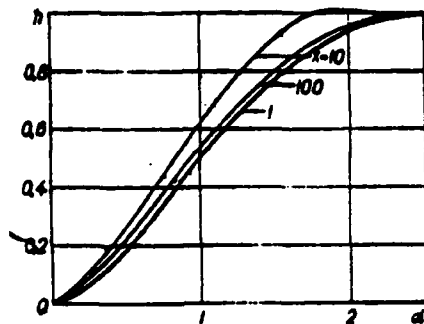


Fig. 90. The transient responses of nonlinear cascade/stage with the optimum parameters.

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§4. Transient processes in selective FU with the nonlinear amplifier instruments.

Selective amplifiers on the tubes.

Tuned amplifier with the single oscillatory circuit.

During the analysis of transient processes in the vacuum-tube amplifiers we will consider that the tube is inertia-free UP. If as UP pentode is used, it is possible to consider that the transient processes, which take place in the cathode circuit, do not depend on transient processes in the anode circuit. Therefore the analysis of

transient processes it is expedient to carry out into two stages: first in the cathode circuit, and then in the anodic.

Furthermore, the mathematical solution of stated problem considerably is facilitated, if transient processes are examined for two input signal levels: for small levels of the signal, when is performed inequality $U_m < U_0 - U_{cm. n}$; for the high levels of the signal, when inequality $U_m > U_0 - U_{cm. n}$ is fulfilled

Transient processes on a small signal level.

Transient processes in the cathode circuit. On a small signal level the transient response of tube can be approximated by the polynomial of the second power. In this case transient processes in the cathode circuit are described by equation (3-36), which can be registered in the following form:

$$\frac{dx}{ax^2 + bx + c} = -\frac{1}{C_n} dt, \quad (5-103)$$

where

$$a_1 = a; \quad a_1 + \frac{1}{R} = b; \quad a_2 - \frac{E}{R} + 0.5a_2 U_m^2 = c'; \quad a_2 - \frac{E}{R} = c;$$

$$\sqrt{b^2 - 4ac} = s; \quad \sqrt{b'^2 - 4ac'} = s'; \quad E - U = x.$$

Equation (5-103) is the equation with that dividing with variables, which easily is led to the quadrature:

$$t = \int \frac{dx}{ax^2 + bx + c} + M. \quad (5-104)$$

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The final solution of integral (5-104) after the determination of integration constant M , on the basis of initial conditions $U_m = 0, x = x_0 = \frac{-b + s}{2a}$ with $t=0$, takes the following form

$$x = -\frac{1}{2a} \left[b - s' \operatorname{cth} \left(\frac{s'}{2C_n} t + \operatorname{Arcth} \frac{s}{s'} \right) \right]. \quad (5-105)$$

Transient processes on a small signal level are described by expression (5-105). When $t = \infty \operatorname{cth} \left(\frac{s'}{2C_n} t + M \right) = 1$ and, therefore,

$$x = x_\gamma = \frac{-b + s'}{2a}. \quad (5-106)$$

Formula (5-106), obtained as a result of solving differential equation (5-103), coincides with previously obtained formula (3-37).

After determining x , we will obtain the law of a change in the amplitude of the fundamental harmonic of the anode current

$$I_{a1} = [a_1 + 2a_2(E - U)] U_m = \left[s' \operatorname{cth} (\alpha t + M) - \frac{1}{R_n} \right], \quad (5-107)$$

where

$$\alpha = \frac{s'}{2C_n}; \quad M = \operatorname{Arcth} \frac{s}{s'}.$$

Transient processes in the anode circuit. For the analysis of transient processes in the anode circuit we will use the equivalent

diagram, depicted in Fig. 91a. Since the cascade/stage is assembled on the parallel diagram of feed, it is expedient to represent duct/contour in the form of three parallel circuits R_0 , L and C_0 . According to the law of Kirchhoff

$$i_a(t) = i_R(t) + i_L(t) + i_C(t).$$

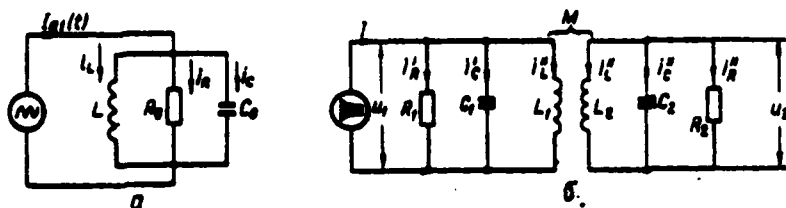


Fig. 91. Equivalent diagrams of the anode circuit of stages: a) resonance; b) band.

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Since

$$i_a(t) = I_{a1}(t) e^{j\omega t} = A(t) e^{j\omega t}; \quad i_R = \frac{u}{R_0}; \quad i_C = C_0 \frac{du}{dt};$$

$$\frac{di_L}{dt} = \frac{u}{L}; \quad \frac{di_R}{dt} = \frac{1}{R_0} \cdot \frac{du}{dt}; \quad \frac{di_C}{dt} = C_0 \frac{d^2u}{dt^2},$$

then finally

$$\frac{d^2u}{dt^2} + 2\delta \frac{du}{dt} + \omega_0^2 u = \frac{1}{C} \cdot \frac{d}{dt} [A(t) e^{j\omega t}]. \quad (5-108)$$

Value $\delta = \frac{1}{2C_0 R_0}$ we will consider small and $\omega = \omega_0$. Then equation can be solved by the method of the slowly varying amplitudes. Let us compose the shortened equation. We will seek solution in the form $u(t) = U(t) e^{j\omega t}$. Finding the derivatives of the unknown solution and substituting in equation (5-108), we obtain

$$\frac{d^2U}{dt^2} e^{j\omega t} + 2j\omega \frac{dU}{dt} e^{j\omega t} - \omega^2 U e^{j\omega t} + 2\delta \left(\frac{dU}{dt} e^{j\omega t} + j\omega U e^{j\omega t} \right) + \omega_0^2 U e^{j\omega t} = \frac{1}{C} \left(\frac{dA}{dt} e^{j\omega t} + j\omega A e^{j\omega t} \right).$$

In accordance with the method of the composition of the shortened equations it is disregarded by value d^2U/dt^2 in comparison with $\omega(dU/dt)$; $\delta(dU/dt)$ in comparison with $\delta\omega U$ and dA/dt in comparison with ωA . Then we obtain the shortened equation:

$$\frac{dU}{dt} + \delta U = \frac{1}{2C_0} A(t), \quad (5-109)$$

or

$$\frac{dU}{dt} + \delta U = f \cosh(\alpha t + M) - g, \quad (5-109a)$$

where

$$f = \frac{U_m \epsilon'}{2C_0}; g = \frac{E}{2C_0 R_n}; \alpha = \frac{\epsilon'}{2C_n}; M = \text{Arcth} \frac{\epsilon'}{\epsilon'}$$

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The obtained equation is nonhomogeneous linear first-order equation

$$\frac{dz}{ds} + Pz = Q,$$

solution of which takes the form

$$z = e^{-\int P ds} [C_1 + \int Q e^{\int P ds} ds].$$

Then for equation (5-109) we obtain

$$U = \frac{1-g}{\delta} + C_1 e^{-\alpha t} + 2f \frac{e^{-\alpha t}}{\delta} \int \frac{ds^{\alpha t}}{e^{\alpha t} s^{2\alpha t} - 1},$$

where $\alpha^2 = \epsilon^2 M$.

In integral $J = \int \frac{ds^{\alpha t}}{e^{\alpha t} s^{2\alpha t} - 1}$ we make substitution $e^{\alpha t} = z$ and we solve it for the series/row of values $\alpha = \delta/4, \delta/2, \delta$ and 2δ . After the substitution of integration constant, found from the initial conditions $U=0$ at $t=0$, and values f and g , finally we obtain:

$$\begin{aligned}
 1. \quad \alpha &= \frac{\delta}{2} \\
 \frac{U(t)}{U_m R_0} &= \left(s' - \frac{1}{R_n}\right)(1 - e^{-st}) + 4s'e^{-2Mt} \left[\left(e^{-\frac{st}{2}} - e^{-st}\right) + \right. \\
 &\quad \left. + e^{-2Mt} e^{-st} \ln \frac{e^{\frac{2M+st}{2}} - 1}{e^{2M} - 1} \right]; \\
 2. \quad \alpha &= \frac{\delta}{2} \\
 \frac{U(t)}{U_m R_0} &= \left(s' - \frac{1}{R_n}\right)(1 - e^{-st}) + 2s'e^{-2Mt} e^{-st} \ln \frac{e^{2M+st} - 1}{e^{2M} - 1}; \\
 3. \quad \alpha &= \delta \\
 \frac{U(t)}{U_m R_0} &= \left(s' - \frac{1}{R_n}\right)(1 - e^{-st}) + 2s'e^{-Mt} e^{-st} (\text{Arcths}^M - \\
 &\quad - \text{Arcths}^{M+st}); \\
 4. \quad \alpha &= 2\delta \\
 \frac{U(t)}{U_m R_0} &= \left(s' - \frac{1}{R_n}\right)(1 - e^{-st}) + s'e^{\frac{M}{2}} e^{-st} [\text{Arcths}^{\frac{M}{2}} - \\
 &\quad - \text{Arcths}^{\frac{M}{2}+st} - (\text{arctg} e^{\frac{M}{2}} - \text{arctg} e^{\frac{M}{2}+st})].
 \end{aligned} \tag{5-110}$$

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From obtained expressions (5-110) it is evident that at the different values α , i.e., with different C_n and U_m ,

the transient processes have different character. From formulas (5-110) it is directly difficult to see, as the character of transient processes with a change in the relationship/ratio α/δ is changed. Therefore it is expedient to consider a numerical example for the concrete/specific/actual diagram and results of calculation

to depict graphically.

For an example let us take cascade/stage UPCh, assembled on the tube of the type 6Zh1P with operational conditions $U_a = 120$ V and $U_g = 100$ V. The characteristic of tube for the selected mode/conditions is approximated sufficiently well by the polynomial of second power [12]

$$i = 20 + 9U + 1U^2.$$

The given approximation it is possible to use in the limits of grid voltage 0-5 V, when polynomial curve coincides well from the tube averaged by characteristic. We choose initial bias voltage $U_{cm. n} = x = E - U = -2.5$ V. Anode current $I_{a. n} = 3.75$ mA. with the selected bias voltage and approximation indicated transient processes in the diagram can be calculated for the signals with the amplitude, which does not exceed $U_m < 1-2.5$ V. We choose resistor/resistance in cathode circuit $R_n = 4$ kilohm. Then $U_n = 3.75 \cdot 10^{-3} \cdot 4 \cdot 10 = 15$ V;
 $E = U - U_{cm. n} = 15 - 2.5 = 12.5$ V.

Let us assume that the cascade/stage has the following parameters: resonance frequency $f_0 = 30$ MHz; passband $\Pi = 2$ MHz; the capacity/capacitance of plate circuit $C_0 = 40$ pF.

Then the total resistance of plate load $R_0 = \frac{1}{2\pi f C_0} = 2$ kilohm.

We determine transient processes with the amplitude of input signal $U_m = 2$ V. Then formulas (5-110) take the form:

$$\frac{U}{U_m R_0} = 2,92(1 - e^{-t'}) + 1,84 \left[(e^{-\frac{t'}{2}} - e^{-t'}) + 0,145 e^{-t'} \ln \frac{0,82 e^{\frac{t'}{2}} - 1}{0,88 - 1} \right];$$

$$\frac{U}{U_m R_0} = 2,92(1 - e^{-t'}) + 0,92 e^{-t'} \ln \frac{0,81 \cdot e^{t'} - 1}{0,58 - 1};$$

$$\frac{U}{U_m R_0} = 2,92(1 - e^{-t'}) + 2,41 e^{-t'} (\text{Arcth } 2,62 - \text{Arcth } 2,62 e^{t'});$$

$$\frac{U}{U_m R_0} = 2,92(1 - e^{-t'}) + 1,95 e^{-t'} [\text{Arcth } 1,62 - \text{Arcth } 1,62 e^{t'} - (\text{arctg } 1,62 - \text{arctg } e^{t'})].$$

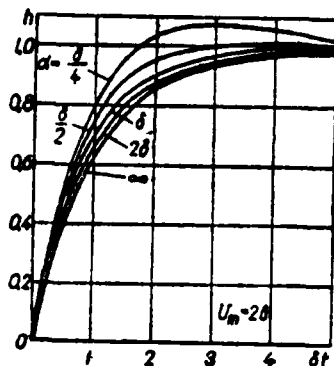


Fig. 92. The transient responses of resonance cascade/stage on a small level of signal $U_m = 2$ V.

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The calculated transient responses of cascade/stage $h = \frac{U}{U_{\text{var}}} = f(\delta t)$ are depicted in Fig. 92. From the figure one can see that, the less the value α , i.e., the greater the value of capacity/capacitance C_m , the less the pulse rise-time. At a certain value α parasitic overshoot appears at the flat/plane pulse apex. Thus, for instance, with $\alpha = \delta/4$ overshoot composes approximately/exemplarily $\Delta = 7\%$. Parasitic overshoot is reduced with the reduction of the amplitude of input signal. After conducting of calculations for $U_m = 1$ V and $\alpha = \delta/4$ it is found with $\Delta = 0.4\%$.

Transient processes on the high signal level.

Transient processes in the cathode circuit. Under the influence on the input of large signal it is expedient to approximate the characteristic of tube by the broken straight line (Fig. 48a). The character of transient processes in anodic circuit is determined by the transient process of changing the voltage/stress U from initial value $U_{n.0}$ at moment $t=0$ of the inclusion/connection of the jump of input voltage to conservative value U_n , (Fig. 93). Therefore, for the solution of the problem sufficiently determining the character of a change in increment U'_n in voltage/stress U_n i.e.

$$U'_n(t) = U_n(t) - U_{n.0}$$

With fulfilling of inequality $U_m > U_0 - U_{m.0}$ the initial total conduction angle of tube θ_n insignificantly differs from value of 90° . In the first approximation, it is possible to consider that $\theta_n = 90$; $U_{m.0} = U_0$; $U_{n.0} = 0$ and the operating point in the initial state with $t=0$ is located on the axis of abscissas.

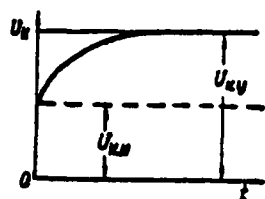


Fig. 93. Transient curves in the cathode circuit of cascade/stage.

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Although this mode/conditions for the low signals is not realistic, it sufficiently close to the real with the large signals. With the adopted assumption the solution of stated problem considerably is simplified, since it is possible to use relationships/ratios known from the theory of diode detection. Error in quantitative estimation is insignificant, which is confirmed by experiment.

With the made assumption the constant component of cathode current and the fundamental harmonic of anode current I_{a1} is respectively equal to:

$$I_{km} = \frac{S_N U_m}{\pi} (\sin \theta - \theta \cos \theta); \quad (5-111)$$

$$I_{a1} = \frac{S_N U_m}{\pi} (\theta - \sin \theta \cos \theta) \quad (5-112)$$

performed
and the equalities

$$U'_N = U_m \cos \theta; \quad (5-113)$$

$$U'_{N\gamma} = U_m \cos \theta_\gamma; \quad (5-114)$$

$$\frac{\tan \theta - \theta}{\pi} = \frac{1}{S_N R_N}. \quad (5-115)$$

Taking into account (3-35) and (5-11), we obtain

$$C_n \frac{dU'_n}{dt} + \frac{U'_n}{R_n} = \frac{S_n U_m}{\pi} (\sin \theta - \theta \cos \theta),$$

since in this case

$$i_R = \frac{U'_n}{R_n}; \quad i_C = C_n \frac{dU'_n}{dt}.$$

After the exception/elimination of value U_m

$$\frac{dU'_n}{dt} + \frac{U'_n}{C_n R_n} = \frac{S_n U_n}{\pi} (\sin \theta - \theta \cos \theta). \quad (5-116)$$

Introducing the relative voltage/stress

$$\text{we obtain} \quad x = \frac{U'_n}{U_{n. \gamma}} = \frac{\cos \theta}{\cos \theta_{\gamma}}, \quad (5-117)$$

$$\frac{dx}{dt} = f(x), \quad (5-118)$$

where

$$f(x) = \frac{S_n}{\pi C_n} (\lg \theta - \theta) x - \frac{x}{C_n R_n}. \quad (5-119)$$

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Differential equation (5-118) is led to the integral of the form

$$t = \int_0^x \frac{dx}{f(x)}, \quad (5-120)$$

which can be computed graphically.

However, the graphic method of integration is very bulky and, using it, it is possible to obtain only the particular solutions, valid only for assigned parameters $R_n - C_n$. Integral (5-120) it is possible to compute analytically, approximating function $f|x|$ by the equation of the parabola of the form

$$F(x) = A(1 - Bx)(1 - x), \quad (5-121)$$

where A and B - the constant coefficients, the methodology of determination of which is given in Appendix 3.

Substituting expression (5-121) into integral (5-120), we obtain

$$At = \int_0^1 \frac{dx}{(1-x)(1-Bx)}. \quad (5-122)$$

Integral (5-122) is tabular

$$At = \frac{1}{1-B} \ln \frac{1-Bx}{1-x}.$$

Whence

$$x = \frac{1 - e^{-\gamma t}}{1 - B e^{-\gamma t}}, \quad (5-123)$$

where

$$\gamma = A(1-B) = f(0)(1-B). \quad (5-124)$$

Reduced coordinates (Fig. 94) shows family of curves, calculated according to formula (5-123), for slope/transconductance $S_n = 4.5$ ma/V and three values of resistor/resistance $R_n = 10$ to kilohm; 4 kilohm; 1 kilohm. Along the axis of abscissas the dimensionless quantity $\alpha = \gamma t$ is deposited/postponed.

From the curves, depicted for Fig. 94, it is possible to design the time of establishment t_p , during which voltage U_n in the cathode attains 90% of conservative value.

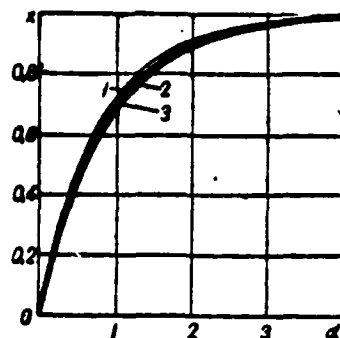


Fig. 94. Family of curves $x=f(\alpha)$ in the reduced coordinates:

$$\begin{aligned} 1 - R_n = 10 \frac{\Omega}{\text{mm}^2}; & \quad 2 - R_n = 4 \frac{\Omega}{\text{mm}^2}; \\ 3 - R_n = 1 \frac{\Omega}{\text{mm}^2}. \end{aligned}$$

Key: (1). kilohm.

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Family of curves $t_y = \varphi(R_n)$, that shows the dependence of the value of time t_y from value R_n for the different values of capacity/capacitance C_n , is given in Fig. 95. From the curves in Fig. 95 evident that with an increase in resistor/resistance R_n time t_y first sharply increases, and then it remains almost constant, i.e., time t_y depends nonlinearly on value R_n . With an increase in capacitance C_n time t_y increases directly proportional to value C_n .

Transient processes in the anode circuit. Calculated curve of relative as a change in the amplitude of the fundamental harmonic of anode current $\beta = \frac{I_{a1}}{I_{a17}} = f(\alpha)$ for $R_n = 4$ kilohm is depicted in Fig. 96.

Calculation is made according to formula (5-123) with the use by curved 2, given in Fig. 94.

The curve, given in Fig. 96, with the high degree of accuracy is approximated by function $\beta = 1 + 0,47e^{-\gamma t}$ in the relative coordinates (point in Fig. 96) or by function $\beta = 1 + 0,47e^{-\gamma t}$ in time coordinates. Value γ is given in Table 11.

Thus, for determining the character of transient processes in the anode circuit of cascade/stage on the high signal level it is necessary to consider effect on the oscillatory circuit of the fundamental harmonic of anode current, which is changed according to the law

$$I_{a1}(t) = (B + De^{-\delta t}) e^{j\omega t}. \quad (5-125)$$

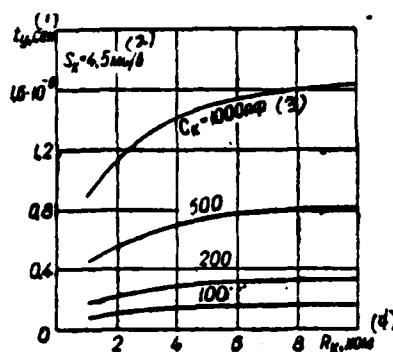


Fig. 95.

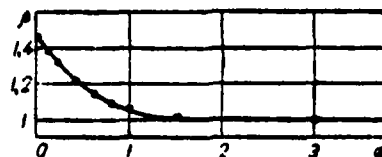


Fig. 96.

Fig. 95. Curves of dependence of time of establishment of video pulse into cathode circuit on value R_n and C_n .

Key: (1). s. (2). ma/V. (3). pF. (4). kilohm.

Fig. 96. Law of change in amplitude of fundamental harmonic of anode current: — $\beta = 1/(1 + 0.47d)$; $\beta = 1 + 0.47e^{-2d}$

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Using a method of the slowly changing amplitudes, we obtain the following shortened equation:

$$\frac{dU}{dt} + \gamma U = \frac{1}{2C_0} (B + D e^{-bt}), \quad (5-128)$$

where $b = 2\gamma$.

Equation (5-126) is linear. It is easy to solve it by operational method. We write/record equation (5-126) in the operational form

$$pU(p) + \delta U(p) = \frac{1}{C_0} \left(B + D \frac{p}{p+b} \right).$$

Whence

$$U(p) = \frac{1}{2C_0} \left[\frac{B}{p+b} + D \frac{p}{(p+b)(p+b)} \right]. \quad (5-127)$$

Through the tables of operational calculus we find original of the image

$$U(t) = \frac{B}{2C_0} (1 - e^{-bt}) + \frac{D(e^{-bt} - e^{-bt})}{2C_0(b-b)}. \quad (5-128)$$

With $b=\delta$ the uncertainty/indeterminacy occurs. Using l'Hopital's rule, we find

$$U_{b=\delta}(t) = \frac{B}{2C_0} (1 - e^{-\delta t}) + \frac{D}{2C_0} t e^{-\delta t}. \quad (5-129)$$

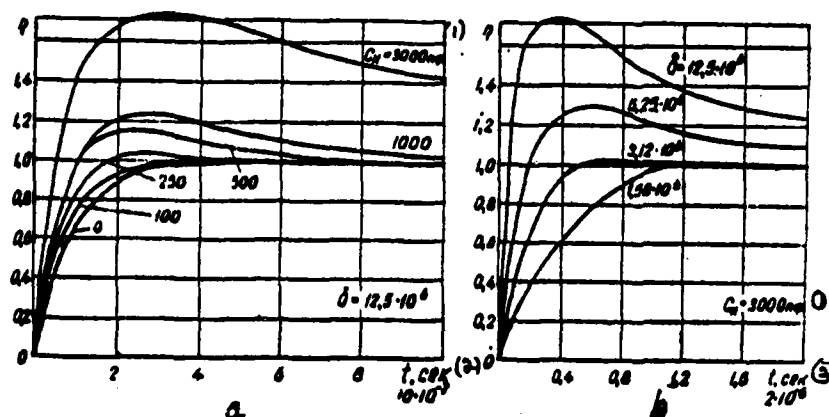


Fig. 97. The transient responses of resonance cascade/stage on the high level of signal $U=6$ V.

Key: (1). pF. (2). s.

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Expressions (5-128), (5-129) to more conveniently represent in the form of transient responses $h = \frac{U(t)}{U_0}$, led to one

$$h = (1 - e^{-bt}) + \frac{D\delta(e^{-at} - e^{-bt})}{B(b-a)}; \quad (5-130)$$

$$h_{\delta=\text{var}} = (1 - e^{-bt}) + \frac{D}{B}\delta t e^{-bt}. \quad (5-131)$$

Calculated transient responses for the resonance cascade/stage with the passband $\Pi=4$ MHz ($\delta=12.6 \cdot 10^{-4}$) at the different values of capacitance C_n are depicted in Fig. 97a, and with $\delta=\text{var}$ and $C_n=3000$ pF = const - 97b. In both cases $R_n=4$ kilohm.

Experimental data coincide sufficiently well with the calculated ones.

Selective amplifier with the two-circuit band-pass filter.

Since the transient processes, which take place in the cathode circuit of band-pass amplifier, have the same character, that for the tuned amplifier, the problem in this case is reduced to the determination of voltage/stress on the coupled circuits with the course through the first plate circuit of the current, which is changed according to the law (5-125). The expressions, which describe transient processes under the influence of low and large signals, are identical. Therefore, let us lead total analysis.

The equivalent diagram of cascade/stage is depicted in Fig. 91b. Let us assume that the first and secondary circuits are identical, i.e., $R_1=R_2=R$, $C_1=C_2=C$. For the diagram, depicted in Fig. 91b we compile an equation of Kirchhoff

$$i = i'_R + i'_C + i'_L; \quad (5-132)$$

$$i''_R + i''_C + i''_L = 0. \quad (5-133)$$

Furthermore,

$$u_1 = i'_R R = \frac{1}{C} \int i'_C dt = L \frac{di'_L}{dt} + M \frac{di''_L}{dt}; \quad (5-134)$$

$$u_2 = i''_R R = \frac{1}{C} \int i''_C dt = L \frac{di''_L}{dt} + M \frac{di'_L}{dt}. \quad (5-135)$$

For the exception/elimination of value i'_R , i'_C , i''_R and i''_C we differentiate equations (5-134) and (5-135) and results let us substitute in differentiated equations (5-132) and (5-133). We will obtain four equations with four unknowns

$$\frac{di'_L}{dt} = \frac{u_1}{L} - \frac{M}{L} \cdot \frac{di''_L}{dt}; \quad (5-136)$$

$$\frac{di''_L}{dt} = \frac{u_2}{L} - \frac{M}{L} \cdot \frac{di'_L}{dt}; \quad (5-137)$$

$$C \frac{d^2 u_1}{dt^2} + \frac{1}{R} \cdot \frac{du_1}{dt} + \frac{u_1}{L} - \frac{M}{L} \cdot \frac{di''_L}{dt} = \frac{di}{dt}; \quad (5-138)$$

$$C \frac{d^2 u_2}{dt^2} + \frac{1}{R} \cdot \frac{du_2}{dt} + \frac{u_2}{L} - \frac{M}{L} \cdot \frac{di'_L}{dt} = 0. \quad (5-139)$$

In order to avoid cumbersome calculations, during the solution of equations (5-136)-(5-139) we apply the following artificial reception/procedure. We store/add up and it is subtracted in pairs equation (5-136) and (5-137), (5-138) and (5-139), and then the results of the actions above equations (5-136) and (5-137) we substitute in the equations, obtained during addition and subtraction of equations (5-138) and (5-139). After which we obtain

$$C \frac{d^2 (u_1 + u_2)}{dt^2} + \frac{1}{R} \cdot \frac{d(u_1 + u_2)}{dt} + \frac{u_1 + u_2}{L + M} = \frac{di}{dt}; \quad (5-140)$$

$$C \frac{d^2 (u_1 - u_2)}{dt^2} + \frac{1}{R} \cdot \frac{d(u_1 - u_2)}{dt} + \frac{u_1 - u_2}{L - M} = \frac{di}{dt}. \quad (5-141)$$

We introduce the designations

$$\frac{1}{2RC} = \delta; \quad \frac{1}{\sqrt{LC}} = \omega_0; \quad \frac{M}{L} = k; \quad \frac{1}{\sqrt{(L+M)C}} = \frac{\omega_0}{\sqrt{1+k}} = \omega_1;$$

$$\frac{1}{\sqrt{(L-M)C}} = \frac{\omega_0}{\sqrt{1-k}} = \omega_2; \quad u_1 + u_2 = x; \quad u_1 - u_2 = y.$$

Then equation (5-140) and (5-141): they take the form

$$\ddot{x} + 2\lambda\dot{x} + \omega_1^2 x = \frac{1}{C} \cdot \frac{di}{dt}; \quad (5-142)$$

$$\ddot{y} + 2\lambda\dot{y} + \omega_2^2 y = \frac{1}{C} \cdot \frac{di}{dt}. \quad (5-143)$$

The obtained equations are the independent equations of the second power, which it is easy to solve.

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After this as will be found the solutions of equations (5-142) and (5-143), unknown edge stresses will be determined according to the formulas

$$u_1 = \frac{x+y}{2}; \quad (5-144)$$

$$u_2 = \frac{x-y}{2}. \quad (5-145)$$

The solution of equations (5-142) and (5-143) with current $i(t)$, determined by formula (5-107), leads to the integrals of the form

$$\int \text{cth } at e^{-t/\tau} e^{i\omega t} dt,$$

which in the elementary functions are not taken. In order to bring solution to the end/lead, let us replace the function of hyperbolic cotangent with exponential function. Let us show that this replacement is possible. The function is given

$$I_m = U_m \left[e' \text{cth} \left(at + \text{Arcth} \frac{a}{e'} \right) - \frac{1}{R_n} \right].$$

Let us replace with its following function

$$I_m = U_m \left[e' + (e - e') e^{eat} - \frac{1}{R_n} \right], \quad (5-146)$$

where ξ - coefficient, which must be determined for the register of functions.

With $t=0$ both functions have the identical values

$$I_m = U_m \left(\epsilon - \frac{1}{R_m} \right).$$

With $t=\infty$ both functions also have the identical values

$$I_m = U_m \left(\epsilon' - \frac{1}{R_m} \right).$$

For the register of functions in the interval $0 < t < \infty$ let us require their equality at certain fixed/recorded moment of time $\alpha t = \beta$

$$U_m \left[\epsilon' \operatorname{cth} \left(\beta + \operatorname{Arcth} \frac{\epsilon}{\epsilon'} \right) - \frac{1}{R_m} \right] = U_m \left[\epsilon' + (\epsilon - \epsilon') e^{-\beta} - \frac{1}{R_m} \right];$$

$$\operatorname{cth} \left(\beta + \operatorname{Arcth} \frac{\epsilon}{\epsilon'} \right) = 1 + \left(\frac{\epsilon}{\epsilon'} - 1 \right) e^{-\beta}.$$

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Since

$$\operatorname{cth}(\varphi + \psi) = \frac{1 + \operatorname{cth} \varphi \operatorname{cth} \psi}{\operatorname{cth} \varphi + \operatorname{cth} \psi},$$

that

$$\left(\frac{\epsilon}{\epsilon'} - 1 \right) e^{-\beta} = \frac{1 + \frac{\epsilon}{\epsilon'} \operatorname{cth} \beta}{\frac{\epsilon}{\epsilon'} + \operatorname{cth} \beta} - 1.$$

Whence

$$\zeta = \frac{1}{\beta} \ln \frac{\operatorname{cth} \beta + \frac{\epsilon}{\epsilon'}}{\operatorname{cth} \beta - 1}.$$

For that designed earlier an example we have: $\epsilon=4.25$; $\epsilon'=3.17$;
 $\epsilon/\epsilon'=1.34$.

Let $\beta=1$. Then

$$\zeta = \ln \frac{1.313 + 1.34}{1.313 - 1} = 2.13.$$

For checking the coincidence let us calculate the functions

$$f_1 = \operatorname{cth} \left(\alpha t + \operatorname{Arcth} \frac{2}{\gamma} \right) = \operatorname{cth} (\alpha t + 0.90);$$

$$f_2 = 1 + \left(\frac{2}{\gamma} - 1 \right) e^{-2.13\alpha t} = 1 + 0.34 e^{-2.13\alpha t}.$$

αt	0	0.5	1	1.5	2	2.5	3	4
f_1	1.34	1.1165	1.0405	1.0147	1.0054	1.0020	1.00078	1.0001
f_2	1.34	1.1165	1.040	1.0139	1.0047	1.0016	1.00058	1.00007

The greatest difference of the values of functions does not exceed 0.08%, which indicates a good coincidence of functions f_1 and f_2 .

Thus, the feeding current

$$i(t) = U_m \left[e' + (e - e') e^{-\theta t} - \frac{1}{R_m} \right] e^{j\omega t} = D(t) e^{j\omega t},$$

where $\theta = 2.13\alpha$; $D(t) = U_m \left[e' + (e - e') e^{-\theta t} - \frac{1}{R_m} \right]$ - slowly changing amplitude.

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We will write the solution of equation (5-142) in the form

$$x = A(t) e^{j\omega t},$$

where $A(t)$ - the slowly changing amplitude of oscillation.

The shortened equation takes the form

$$2 \frac{dA}{dt} j\omega_0 + A(\omega_1^2 - \omega_0^2) + 2\delta A j\omega_0 = \frac{1}{C} D j\omega_0.$$

Taking into account that $\omega_1^2 - \omega_0^2 = \omega_0^2((1/1+k)-1) = -k\omega_0^2$, we obtain

$$\frac{dA}{dt} + (\delta + j\Omega) A = M + N e^{-st}, \quad (5-147)$$

where

$$\Omega = \frac{\lambda \omega_0}{2}; \quad M = \frac{U_m \left(s' - \frac{1}{R_n} \right)}{2C}; \quad N = \frac{U_m (s - s')}{2C}.$$

This is linear the differential equation, analogous to equation (5-126).

The resultant expression for amplitude $A(t)$, found as a result of solving equation (5-147) taking into account integration constant, takes the following form:

$$A = \frac{M}{\delta + j\Omega} [1 - e^{-(\delta + j\Omega)t}] + \frac{N}{\delta - \delta + j\Omega} [e^{-st} - e^{-(\delta + j\Omega)t}]. \quad (5-148)$$

We analogously find the solution of equation (5-143)

$$y = B(t) e^{j\omega_0 t},$$

$$B = \frac{M}{\delta - j\Omega} [1 - e^{-(\delta + j\Omega)t}] + \frac{N}{\delta - \delta + j\Omega} [e^{-st} - e^{-(\delta + j\Omega)t}]. \quad (5-149)$$

We find the complex amplitudes of edge stresses

$$\begin{aligned} \dot{U}_1 &= \frac{A+B}{2} = \frac{1}{2} \left\{ \frac{M}{\delta + j\Omega} [1 - e^{-(\delta + j\Omega)t}] + \frac{M}{\delta - j\Omega} [1 - e^{-(\delta - j\Omega)t}] + \right. \\ &\quad \left. + \frac{N}{\delta - \delta + j\Omega} [e^{-st} - e^{-(\delta + j\Omega)t}] + \frac{N}{\delta - \delta + j\Omega} [e^{-st} - e^{-(\delta - j\Omega)t}] \right\}; \\ \dot{U}_2 &= \frac{A-B}{2} = \frac{1}{2} \left\{ \frac{M}{\delta + j\Omega} [1 - e^{-(\delta + j\Omega)t}] - \frac{M}{\delta - j\Omega} [1 - e^{-(\delta - j\Omega)t}] + \right. \\ &\quad \left. + \frac{N}{\delta - \delta + j\Omega} [e^{-st} - e^{-(\delta + j\Omega)t}] - \frac{N}{\delta - \delta - j\Omega} [e^{-st} - e^{-(\delta - j\Omega)t}] \right\}. \end{aligned}$$

After multiplying the complex amplitudes \dot{U}_1 and \dot{U} , on $e^{j\omega_0 t}$ and after taking real parts, let us find the instantaneous values of edge stresses

$$u_1 = \left[\frac{Mb}{b^2 + \Omega^2} \left[1 - e^{-bt} \left(\cos \Omega t - \frac{\Omega}{b} \sin \Omega t \right) \right] + \frac{N(b-b)}{(b-b)^2 + \Omega^2} \left[e^{-bt} - e^{-bt} \left(\cos \Omega t - \frac{\Omega}{b-b} \sin \Omega t \right) \right] \right] \cos \omega_0 t; \quad (5-150)$$

$$u_2 = \left[\frac{M\Omega}{b^2 + \Omega^2} \left[1 - e^{-bt} \left(\cos \Omega t + \frac{b}{\Omega} \sin \Omega t \right) \right] + \frac{N\Omega}{(b-b)^2 + \Omega^2} \left[e^{-bt} - e^{-bt} \left(\cos \Omega t + \frac{b-b}{\Omega} \sin \Omega t \right) \right] \right] \sin \omega_0 t. \quad (5-151)$$

Expressions (5-150), (5-151) can be registered in the following form:

$$u_1 = U_1(t) \cos \omega_0 t; \quad (5-152)$$

$$u_2 = U_2(t) \sin \omega_0 t, \quad (5-153)$$

where $U_1(t)$ and $U_2(t)$ - the slowly varying in the time amplitudes of voltages/stresses u_1 and u_2 .

Under the influence of large signal anode current I_{a1} is described by expression (5-125), which it is analogous with function (5-146). Consequently, differential equation (5-147) and results of its solution (5-150) and (5-151) with the substitution of the corresponding values N and M are accurate for the high signal level.

Of greatest interest for the practice is amplitude $U_1(t)$ the voltage/stress of radio pulse, which was isolated on the secondary circuit.

The calculated transient responses h for amplitude $U_1(t)$ are given in Fig. 98a, b

$$h = \frac{U_1(t)}{U_{2v}}, \quad (5-154)$$

where U_{2v} - steady amplitude $U_1(t)$.

For the comparison of the results of research of transient processes in band UPCh with the results, obtained above, the calculations of transient responses are carried out for the cascade/stage, assembled on the tube 6Zh1P, with the cathode resistor of tube $R_k = 4$ kilohm.

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Analyzing Figs 97 and 98, it is possible to do the following conclusions:

at the constant value δ (at the constant value of the passband of cascade/stage) with an increase in capacitance C_n the time of the establishment of the intensive radio pulse is reduced, and the value of parasitic overshoot at the flat/plane pulse apex sharply increases;

at the constant value of capacitance C_n with an increase in the value δ (with the expansion of the passband of cascade/stage) time

δ is reduced, and the value of overshoot increases:

at the identical values of capacitance C_n and δ the parasitic overshoot, which is formed during the amplification of radio pulse in the band cascade/stage, is considerably more than during the amplification in the resonance cascade/stage with the single duct/contour;

both in the resonance ones and in the band cascades/stages, which work on the high levels of signal $U_m > |U_0| - |U_{0.1}|$, the value of capacitance C_n must be chosen in the limits of 100-200 pF. In this case is reached considerable operating speed of MARU and the correction of the front of the intensive radio pulse with a comparatively insignificant increase of parasitic overshoot.

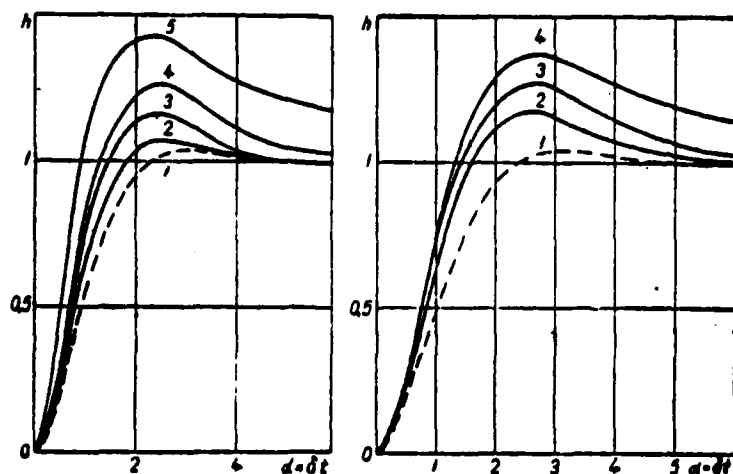


Fig. 98. The transient responses of band cascade/stage with input signal $U_m = 2$ V: a) $\delta = 6.28 \cdot 10^4 = \text{const}$; $c_n = \text{var}$; 1 - $c_n = 0$; 2 - $c_n = 200$ pF; 3 - $c_n = 500$ pF; 4 - $c_n = 1000$ pF; 5 - $c_n = 3000$ pF; b) $c_n = 1000$ pF; $\delta = \text{var}$; 1 - $\delta = 6.28 \cdot 10^4$; $c_n = 0$; 2 - $\delta = 3.14 \cdot 10^4$; 3 - $\delta = 6.28 \cdot 10^4$; 4 - $\delta = 12.56 \cdot 10^4$.

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SELECTIVE TRANSISTORIZED AMPLIFIERS.

Tuned amplifier with the single oscillatory circuit.

The special features/peculiarities of transient processes in the transistorized amplifiers are caused by the form of the passage characteristic of transistor, which with a sufficient degree of accuracy is approximated by exponential curve, and by the inertness

of transistor. Due to the inertness of transistor with an increase in the frequency the slope/transconductance of passage characteristic is reduced, and, therefore, are reduced constant component and fundamental harmonic of collector (emitter) current, and the phase of the reinforced vibrations also is changed.

Since us interests the envelope high-frequency oscillations, a change in the phase can be disregarded/neglected. A decrease in mutual conductance of transistor can be taken into account with the multiplication of the amplitude of the collector current by the frequency factor $m(2-95)$.

In the transistor amplifier, as in the electron-tube, transient processes it is expedient to first consider in the emitter circuit (in the common-emitter connection), and then in the collector. The possibility of separate analysis is caused by the fact that output characteristics of transistor according to the character are similar to the plate characteristics of pentode. During the selection of the corresponding mode/conditions of the work of transistor it is possible to consider that the transient processes in the emitter circuit (in common-emitter connection, Fig. 54c) little depend on transient processes in the circuit of collector/receptacle.

Under the influence on the input of the transistor cascade/stage

of voltage surge of the radio frequency (we consider that is the source of signal the voltage generator with an internal resistor/resistance of $R_e = 0$) in the emitter circuit appears the jump of the constant component of emitter current $I_{e.s}(t)$ with rise time $t_e \approx \frac{T}{4}$, caused by the inertness of transistor, approximately/exemplarily, equal to one fourth of period T of high-frequency oscillations. In comparison with the time of a change in the envelope high-frequency oscillations $t_e = 0$ is possible to take and to consider that in the emitter circuit the instantaneous jump of current $I_{e.s}$ occurs whose value is determined by expression (3-49).

For the emitter circuit of diagram in Fig. 54c it is possible to register the following equation of Kirchhoff:

$$I_{e.s}(t) = I_R(t) + I_e(t). \quad (5-155)$$

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If for $I_{e.s}$ we take expression (3-49), then the equation, which describes transient processes in the emitter circuit, is transcendental and can be solved only graphically. For the analysis of transient processes we use an approximation for $I_{e.s}$ taking into account expression (3-55) .

$$I_{e.s} = \frac{1}{0.8R_e} (\gamma E + 0.68 + \ln x - \gamma U_e), \quad (5-156)$$

where $x = A_e R_e I_e (\gamma U_m)$; ($R_e = R$ in Fig. 54c); $U_e = U$.

Then equation (5-155) can be registered

$$\frac{dU_e}{dt} + \frac{U_e}{R_e} = I_{e, n},$$

or

$$\frac{dU_e}{dt} + \alpha_e U_e = \alpha_e \frac{\gamma E + 0,68 + \ln z}{\gamma + 0,6}, \quad (5-157)$$

where

$$\alpha_e = \frac{1 + \frac{\gamma}{0,6}}{R_e C_e} = \frac{1 + 1,66 \gamma}{R_e C_e}.$$

Equation (5-157), which describes transient processes in the emitter circuit, is linear. The solution of equation (5-157) after the substitution of integration constant takes the form:

$$U_e = \frac{\gamma E + 0,68 + \ln z}{\gamma + 0,6} - e^{-\alpha_e t} \frac{\ln I_e(\gamma U_m)}{\gamma + 0,6}. \quad (5-158)$$

In the beginning of the effect of input voltage with $t=0$

$$U_{e, n} = \frac{\gamma E + 0,68 + \ln z_e}{\gamma + 0,6}, \quad (5-159)$$

where

$$z_e = A_e R_e.$$

In the steady-state mode/conditions with $t \rightarrow \infty$

$$U_{e, \gamma} = \frac{\gamma E + 0,68 + \ln z}{\gamma + 0,6}. \quad (5-160)$$

On the basis of expressions (5-158) - (5-160) it is possible to record

$$U_e = U_{e, \gamma} - e^{-\alpha_e t} (U_{e, \gamma} - U_{e, n}), \quad (5-161)$$

or

$$h_e = \frac{U_e - U_{e, n}}{U_{e, \gamma} - U_{e, n}} = (1 - e^{-\alpha_e t}). \quad (5-162)$$

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Thus, the constant voltage U_e in the emitter circuit of

transistor under the influence of the jump of radio frequency voltage is changed in the time exponentially with the time constant $\tau_0 = \frac{1}{\omega_0}$, determined by the values of elements/cells R_0 and C_0 .

Similarly, it is possible to find the law of a change in the bias voltage of the transistor

or
$$U_{a.e}(t) = \frac{0,68 - 0,68 - \ln x}{1 + 0,6} + e^{-\omega_0 t} \frac{\ln I_0(\gamma U_m)}{1 + 0,6}, \quad (5-163)$$

$$U_{a.e}(t) = U_{a.e.7} + e^{-\omega_0 t} (U_{a.e.2} - U_{a.e.7}). \quad (5-164)$$

Taking into account expression (3-50) for the fundamental harmonic of collector current, we write/record the differential equation, which describes transient processes in the collector circuit

$$\frac{d^2 u_c}{dt^2} + 2\beta \frac{du_c}{dt} + \omega_0^2 u_c = \frac{1}{C_0} \cdot \frac{d}{dt} [N e^{j\omega t}], \quad (5-165)$$

where

$$N = A_{\pi e1} U_{a.e.2} I_1(\gamma U_m).$$

The shortened equation takes the form:

$$\frac{dU}{dt} + \beta U = \frac{1}{2C_0} N. \quad (5-166)$$

Solution of this equation

$$U = e^{-\beta t} \left[C + \frac{A}{2C_0} \int \exp(\zeta e^{-\omega_0 t}) e^{j\omega t} dt \right], \quad (5-167)$$

where $A = A_{\pi e1} I_1(\gamma U_m) \exp[\mu(0,68 - 0,68 - \ln x)];$

$$\zeta = \mu \ln I_0(\gamma U_m); \quad \mu = \frac{1}{1 + 0,6}.$$

If we into the integral, entering expression (5-167), substitute

$y = e^{-\zeta t}$, the latter is converted into the tabular integral for integral ratios $n = \frac{b}{a_0} = 1, 2, 3, \dots$

$$J = \int \exp(\zeta e^{-\zeta t}) e^{at} dt = -\frac{1}{a_0} \int \frac{\exp(\zeta y) dy}{y^{n+1}} = -\frac{1}{a_0} \left[e^{\zeta y} \sum_{k=1}^n \frac{\zeta^{k-1}}{(n-1)(n-2) \dots (n-k) e^{\zeta y - k}} - \frac{\zeta^n}{n!} Ei(\zeta y) \right], \quad (5-168)$$

where $Ei(\zeta y)$ - tabulated exponential integral.

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Let us give the series/row of particular solutions for several values of n .

1. $n=1$ ($\delta = a_0$)

$$U(t) = \frac{A}{2C_0 a_0} e^{-at} \{ \zeta [Ei(\zeta) - Ei(\zeta y)] + e^{at} \exp(\zeta y) - \exp \zeta \}. \quad (5-169)$$

With $t \rightarrow \infty$ conservative value of the voltage/stress

$$U_\infty = \frac{A}{2C_0 a_0} (1 - e^\zeta). \quad (5-170)$$

Then the transient response

$$h = \frac{U(t)}{U_\infty} = e^{-at} (1 - e^\zeta)^{-1} \{ \zeta [Ei(\zeta) - Ei(\zeta y)] + e^{at} e^{\zeta y} - e^\zeta \}. \quad (5-171)$$

2. $n=2$ ($a_0 = \frac{b}{2}$)

$$U(t) = \frac{A}{4C_0 a_0} e^{-at} \{ \zeta^2 [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} (y^{-2} + \zeta y^{-1}) - e^\zeta (1 + \zeta) \}. \quad (5-172)$$

With $t \rightarrow \infty$

$$U_\infty = \frac{A}{4C_0 a_0} [1 - e^\zeta (1 + \zeta)].$$

Then

$$h(t) = e^{-at} [1 - e^\zeta (1 + \zeta)]^{-1} \{ \zeta^2 [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} (y^{-2} + \zeta y^{-1}) - e^\zeta (1 + \zeta) \}. \quad (5-173)$$

$$3. n=3 \left(a_0 = \frac{b}{3} \right)$$

$$U(t) = \frac{A}{12C_0 a_0} e^{-\lambda t} \{ \zeta^3 [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} (y^{-3} + \zeta y^{-2} + \zeta^2 y^{-1}) - e^{\zeta} (1 + \zeta + \zeta^2) \}. \quad (5-174)$$

With $t \rightarrow \infty$

$$U_T = \frac{A}{12C_0 a_0} [1 - e^{\zeta} (1 + \zeta + \zeta^2)];$$

$$h(t) = e^{-\lambda t} [1 - e^{\zeta} (1 + \zeta + \zeta^2)]^{-1} \{ \zeta^3 [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} (y^{-3} + \zeta y^{-2} + \zeta^2 y^{-1}) - e^{\zeta} (1 + \zeta + \zeta^2) \}; \quad (5-175)$$

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$$4. n=m \left(a_0 = \frac{b}{m} \right)$$

$$U(t) = \frac{A}{2mC_0 a_0} e^{-\lambda t} \{ \zeta^m [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} \sum_{i=1}^m \zeta^{m-i} y^{-i} - e^{\zeta} \sum_{i=0}^{m-1} \zeta^i \}. \quad (5-176)$$

With $t \rightarrow \infty$

$$U_T = \frac{A}{2mC_0 a_0} [1 - e^{\zeta} \sum_{i=0}^{m-1} \zeta^i];$$

$$h(t) = e^{-\lambda t} [1 - e^{\zeta} \sum_{i=0}^{m-1} \zeta^i]^{-1} \{ \zeta^m [Ei(\zeta) - Ei(\zeta y)] + e^{\zeta y} \sum_{i=1}^m \zeta^{m-i} y^{-i} - e^{\zeta} \sum_{i=0}^{m-1} \zeta^i \}. \quad (5-177)$$

With the help of the resulting expressions it is possible to sufficiently accurately calculate transient responses for different input signal level. However, these expressions are somewhat bulky and it is expedient to use them for a precise calculation of transient responses in the electronic computers. For the rough estimate of the character of transient processes it is possible to use approximate solution, which we find as follows. We write/record expression

(5-167) in the following form

$$U = e^{-\mu t} \left[C + \frac{A}{2C_0} \int e^{C_0 t} dt \right]. \quad (5-178)$$

We expand function $e^{C_0 t}$ in series/row and are limited by two members. After determination and substitution of integration constant

$$U = AR_0 \left[1 - e^{-\mu t} + \frac{\zeta}{1-\chi} (\mu e^{-\mu t} - e^{-\mu t}) \right], \quad (5-179)$$

where

$$\chi = \frac{C_0}{\gamma}.$$

Since $\gamma \gg 1$, it is possible to take $\mu=1$. Let us consider several particular solutions for the series/row of values χ .

1. $\chi=0$

$$U(t) = AR_0(1 - e^{-t})(\zeta + 1). \quad (5-180)$$

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With $t \rightarrow \infty$

$$U_\gamma = AR_0(\zeta + 1).$$

Then

$$h(t) = (1 - e^{-t}). \quad (5-181)$$

2. $0 < \chi < 1$. For the present instance formula (5-179) is real. With $t \rightarrow \infty$

$$U_\gamma = AR_0.$$

Then

$$h(t) = 1 - e^{-t} + \frac{\zeta}{1-\chi} (e^{-\chi t} - e^{-t}). \quad (5-182)$$

3. $\chi=1$. In this case is obtained the uncertainty/indeterminacy

of form %, pointing out which according to l'Hopital's rule, we obtain

$$h(t) = 1 - e^{-t} + \zeta e^{-t} t. \quad (5-183)$$

4. $\chi > 1$

$$h(t) = 1 - e^{-t} \frac{\zeta}{\chi - 1} (e^{-\chi t} - e^{-t}). \quad (5-184)$$

The transient responses, calculated by formulas (5-181)-(5-183), have much the same character as transient responses for the vacuum-tube amplifier. Difference is in the fact that the overshoot at the flat/plane pulse apex appears on the lower level of the input voltage of signal. The most optimum relationship/ratio is $\alpha_0 = \frac{3}{4}$, with which the adjustment is realized virtually on the envelope of impulse/momentum/pulse and overshoot in entire dynamic range up to the entrance of transistor into the mode/conditions of limitation does not exceed 15-20%.

In this case

$$C_{a, \text{opt}} = \frac{8\alpha}{\pi} (1 + 1.66\gamma).$$

For the band-pass amplifier with the two-circuit filter the transient responses are analogous to the transient responses of vacuum-tube amplifier with the same difference, as for the tuned amplifier.

Inertial UP.

Due to the inertial properties of transistor with an increase in the frequency an increment in the constant component of emitter current and amplitude of the first harmonic of collector current is reduced. Because of this the character of transient processes it is changed with an increase in the frequency of the reinforced signals.

On the basis of materials Chapter 3 it is possible to demonstrate that the voltage/stress between the emitter and the base taking into account the inertial properties of the transistor

$$U_{e,0} = \frac{1}{\gamma} \ln \frac{U_{e,0} + \frac{1}{m} \Delta U_e}{A_e R_e J_e (\gamma U_m)}. \quad (5-185)$$

For the exception/elimination of transcendence we use approximate solution (5-156). Then the dc current component of the emitter

$$I_{e,0} = \frac{\gamma E + \ln s + 0,68 - 0,8 (U_e - U_{e,0}) \left(\frac{1}{m} - 1 \right) - \gamma U_e}{0,6 R_e}. \quad (5-186)$$

Substituting expression (5-186) in (5-155) and solving obtained equation $U_e = (t)$, we obtain the following expression, which describes transient processes in the emitter circuit

$$U_e(t) = U_e - e^{-\gamma t} \Delta U_e. \quad (5-187)$$

where

$$\left. \begin{aligned} U_0 &= \frac{\gamma E + 0,68 + \ln A_0 R_0 I_0 (\gamma U_m) + 0,6 U_{0.2} \left(\frac{1}{m} - 1 \right)}{\frac{0,6}{m} + \gamma}; \\ a_0 &= \frac{\frac{1}{m} + \frac{\gamma}{0,6}}{R_0 C_0}; \quad \Delta U_0 = \frac{1}{\frac{0,6}{m} + \gamma} \ln I_0 (\gamma U_m). \end{aligned} \right\} \quad (5-188)$$

For determining the expression, which describes transient processes in the collector circuit, we will use solution (5-167). Taking into account equation (5-186), after conversions we obtain the following expression for the transient response

$$h(t) = (1 - e^{-a_0 t}) + \zeta e^{-a_0 t} \frac{1 - \exp(a_0 - \beta)t}{1 - z}, \quad (5-189)$$

where

$$\zeta = \frac{\gamma}{\frac{0,6}{m} + \gamma} \ln I_0 (\gamma U_m); \quad z = \frac{a_0}{\beta} = \frac{2R_0 C_0}{R_0 C_0} \left(\frac{1}{m} + \frac{\gamma}{0,6} \right).$$

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On the basis of calculations and experimental research it is established/installed, that the inertial properties of transistor differently affect the character of the transient processes, which take place in the emitter and collector circuits,. The inertness of transistor in the larger measure affects transient processes in the collector and in smaller - to the transient processes in the emitter circuit. From expressions (5-188) it is evident that with fulfilling of inequality $\gamma \gg 0.6/m$ by the indicated effect on the transient processes in the emitter circuit it is possible to disregard.

With an increase in the signal frequency coefficient m is reduced. Because of this the time of the establishment of transient response increases and parasitic overshoot is reduced at the apex/vertex of radio pulse.

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Chapter 6.

Calculation, tuning [REDACTED] adjustment of functional amplifiers.

§ 1. Calculation of logarithmic amplifier with the nonlinear elements/cells.

In the present chapter are given a methodology and examples of the calculation of the most widely used types of functional amplifiers - the logarithmic amplifiers, made on the tubes and the transistors. Other types of functional amplifiers can be calculated, using a material of the previous chapters and the given methodologies.

The order of calculation of functional amplifiers does not depend on the type of amplifier instrument. Therefore the methodologies given below are general/common/total for the amplifiers on the tubes and on the transistors.

For calculating the amplifier with FAKh must be assigned the fundamental parameters, enumerated in § 3 of Chapter 1. Amplifier independent of type and method of its realization, they calculate into two stages:

1. Selection of diagram and the calculation of amplifier in the linear conditions according to the given values of the factor of amplification K_0 , passband Π and frequency of the reinforced f_0 . As a result of calculation they must be determined: a quantity of cascades/stages n , the factor of amplification of one cascade/stage K_n , noise voltage on the output of amplifier and value of the network elements, which ensure linear conditions of the work of amplifier.

2. Calculation of network elements, which ensure realization of FAKh in amplifier.

The order of calculation of aperiodic, resonance and band-pass amplifiers to LAKh is identical. The calculation of tuned amplifier is performed employing the following procedure.

1. Input and output voltage of cascade/stage is determined, with which it must begin with its LAKh,

$$u_{\text{NL}, n} = U_{\text{NL}, n} K_n^{n-1}; \quad (8-1)$$

$$u_{\text{NL}, n} = u_{\text{NL}, n} K_n = U_{\text{NL}, n} K_n^n \quad (8-2)$$

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2. Required number of nonlinear cascades/stages is determined. if coefficient K , considerably exceeds dynamic range, then there is no need for making all cascades/stages with nonlinear ones. Quantity of nonlinear cascades/stages with the successive work taking into account condition (1-76), obviously,

$$n_{\text{NL}} = \frac{D_{(00)}}{d_{(00)}} = \frac{D_{(00)}}{K_n(00)}. \quad (8-3)$$

3. Basis of logarithmic signal N (or coefficient $a = \frac{1}{\ln N}$) is determined.

If the dynamic ranges D and D_{NL} , are assigned then according to expression (1-44)

$$a = \frac{D_{\text{NL}} - 1}{\ln D}. \quad (8-4)$$

If slope/transconductance σ and voltage/stress $U_{\text{NL}, n}$, is assigned then according to formula (1-42)

$$a = \frac{\sigma \left[\frac{\sigma}{N \sigma n} \right]}{U_{\text{NL}, n}}. \quad (8-5)$$

If $a=1$, all nonlinear cascades/stages are performed by identical ones. If $a_{\text{NL}} \neq 1$, in the latter/last cascade/stage it is necessary to

ensure amplitude characteristic with $a_{\text{pac}} \neq 1$, and in all rest, which precede the latter, mode/conditions with $a=1$.

4. If amplifier instrument is linear, then according to formula (2-115) or (2-116) taking into account (1-30) and (1-31) is calculated required dependence of change in conductivity $g_{\text{non. o}} = (U_{\text{non. o}})$ or resisting $R_{\text{non. o}} = f(U_{\text{non. o}})$ nonlinear element/cell. In this case, the calculation is performed for the linear conditions with the change $0 \leq z \leq 1$, logarithmic - when $1 < z < d = K_n$, quasi-linear - when $d < z < z_{\text{nt}} = la \ln d$, where i - the reference number of nonlinear cascade/stage.

If amplifier instrument nonlinear, it is possible to calculate dependence $g_{\text{non. o}} = f(U_{\text{non. o}})$, using formula (2-126)

$$g_{\text{non. o}} = g_n \left[\frac{S}{S_n} \varphi(z) - 1 \right] = g_n [\Phi(S) \varphi(z) - 1], \quad (6-8)$$

or volt-ampere characteristic of nonlinear element/cell, using formula (2-127).

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5. Is chosen nonlinear element/cell, usually semiconductor diode, which satisfies following requirements: slope/transconductance of volt-ampere characteristic of diode must be large with small

voltages on diode; stray capacitance of diode and volumetric strength of materials of diode must be small.

6. For different values of cutoff voltage for diode U_c from volt-ampere characteristic of diode by method of five or more than ordinates [36] is designed family of real curves of dependence of conductivity (resistor/resistance) of diode on applied to it sine voltage $g_{\text{non. p}} = f(U_m)$.

7. Curves $g_{\text{non. o}} = f(U_{\text{non. o}})$ and $g_{\text{non. p}} = \varphi(U_m)$ are combined. If required curve $g_{\text{non. o}} = f(U_{\text{non. o}})$ coincides not with one of curves $g_{\text{non. p}} = \varphi(U_m)$, is adapted/hurried one of nearest curves $g_{\text{non. p}} = \varphi(U_m)$ by an increase in the number of in parallel connected nonlinear elements/cells and by inclusion/connection consecutively/serially or in parallel with the nonlinear element of active linear resistor/resistance.

Example of calculation. To calculate logarithmic UPCh with the single resonant circuits, if are assigned the following technical specifications: passband $\Pi = 1$ MHz; amplification factor in linear conditions $K_s = 10^4$; resonance frequency $f_s = 30$ MHz; dynamic range of LARh $D = 80$ dB; the slope/transconductance of LARh $\sigma = 0.2$ V/Np;

$$U_{\text{sat. u}} \approx 2 \cdot 10^6 \text{ V.}$$

The 1st stage. We choose a tube of the type 6Zh5P. Let us assume

that as a result of calculating the amplifier in the linear conditions are acquired following data: $n=4$; $K_H=10$.

The 2nd stage.

1. We determine voltages/stresses $u_{\text{ex. n}}$ and $u_{\text{max. n}}$:

$$u_{\text{ex. n}} = u_{\text{ex. n}} K_H^{n-1} = 2 \cdot 10^{-3} \text{ V};$$

$$u_{\text{max. n}} = u_{\text{ex. n}} K_H = 2 \cdot 10^{-3} \cdot 10 = 0,2 \text{ V}.$$

Key: (1). V.

2. We determine number of nonlinear cascades/stages:

$$n_{\text{max}} = \frac{D_{(\text{sd})}}{K_H(\text{sd})} = \frac{80}{20} = 4.$$

3. We determine coefficient of a:

$$a = \frac{u}{u_{\text{max. n}}} = \frac{0,2}{0,2} = 1.$$

Thus, all nonlinear cascades/stages must be identical.

4. We determine maximum input voltage on input of latter/last (4th) nonlinear cascade/stage

$$u_{\text{ex. max}} = K_H u_{\text{ex. n}} [(1 - a) \ln a + 1] = 10 \cdot 0,02 [(4 - 1) \ln 10 + 1] = 1,58 \text{ V}.$$

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During the appropriate selection of the mode of operation of

tube it is possible to count its characteristic of linear.

5. We choose the method of shunting of plate load of cascade/stage by nonlinear element/cell (Fig. 99). According to formula (2-16) taking into account (1-30) and (1-31) we calculate and plot a curve $R_{\text{нел. с}} = f(U_{\text{нел. с}})$ for the fourth nonlinear cascade/stage. The designed curve as prime is depicted in Fig. 100. For the remaining nonlinear cascades/stages curves $R_{\text{нел. с}} = f(U_{\text{нел. с}})$ will be the same, but finished with the voltages/stresses

$$U_{\text{нел. с. 4}} = K_{\text{н}} U_{\text{нел. с. 1}} (i \ln d + 1).$$

6. As nonlinear element/cell we choose semiconductor diode of type D2Zh. From the method of five ordinates we calculate real curves $R_{\text{нел. д}} = \varphi(U_{\text{н}})$ for the different values of cutoff voltage $U_{\text{с}}$ on the diodes and we represent as unbroken curves in Fig. 100. From this figure it is evident that the curve of the required law of a change of resisting the nonlinear element/cell $R_{\text{нел. с}} = f(U_{\text{нел. с}})$ sufficiently it coincides precisely with the curve of a true change of resisting the nonlinear element/cell $R_{\text{нел. с}} = \varphi(U_{\text{н}})$ when $U_{\text{с}} = 0.25$ V in the range of voltages/stresses 0.25-0.8 V, and then these curves diverge. The disagreement of curves can be reduced by the inclusion/connection consecutively/serially with the nonlinear element of supplementary linear resistor/resistance $R_{\text{л}} = 100$ ohm, which is equal to a difference in resistors/resistances $R_{\text{нел. с}} = 218$ ohm and $R_{\text{нел. д}} = 100$ ohm when $U_{\text{с}} = 0.25$ V at point $U_{\text{нел. с}} = 2.5$ V.

It is expedient to take supplementary resistors/resistances R_d for different nonlinear cascades/stages by different ones. The values of these resistors/resistances it is necessary to find as a difference in resistors/resistances $R_{\text{нел. } p}$ and $R_{\text{нел. } o}$, found from curves $R_{\text{нел. } o} = f(U_{\text{нел. } o})$ and $R_{\text{нел. } p} = \varphi(U_m)$ at voltages/stresses $U_{\text{нел. } o}$ of those corresponding to the end/lead of LAKh of amplifier. In this case voltages/stresses $U_{\text{нел. } o}$ with which it is necessary to find values of R_d for different nonlinear cascades/stages, are given in Table 12. The values of supplementary resistors/resistances, found with the method indicated, are given in Table 13, from which it is evident that for the first nonlinear cascade/stage resistor/resistance $R_d = 0$, and for the rest - it virtually one and the same.

Table 12.

	1	2	3	4
$U_{max} \cdot 10^3$	0,66	1,12	1,58	2,04

Key: (1). V.

Table 13.

	1	2	3	4
$R_{nog. 0.25}$	0	87	90	94

Key: (1). ohm.

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Supplementary resistors/resistances it is expedient to use for the creation on them of those locking nonlinear elements of voltages/stresses U_* .

The points of the passage of the curve of a change of resisting the nonlinear element/cell with a series-connected supplementary resistor/resistance of $R_d = 100$ ohm are shown in Fig. 100. The points of curve $R_{nog. p} + 100 \text{ ohm} = \varphi(U_m)$ coincide sufficiently well with required curve $R_{nog. o} = f(U_{nog. o})$ in entire range of the output voltage/stress of latter/last nonlinear cascade/stage. Because of this it is possible to obtain AKh cascade/stage, the ensuring successive work of all five cascades/stages and precise LAKh of UPCh.

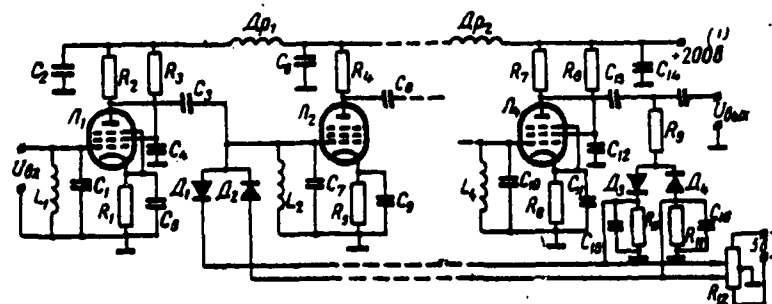


Fig. 99. Schematic diagram of logarithmic tuned amplifier with the nonlinear elements/cells in the anode circuits of cascades/stages.

Key: (1). V.

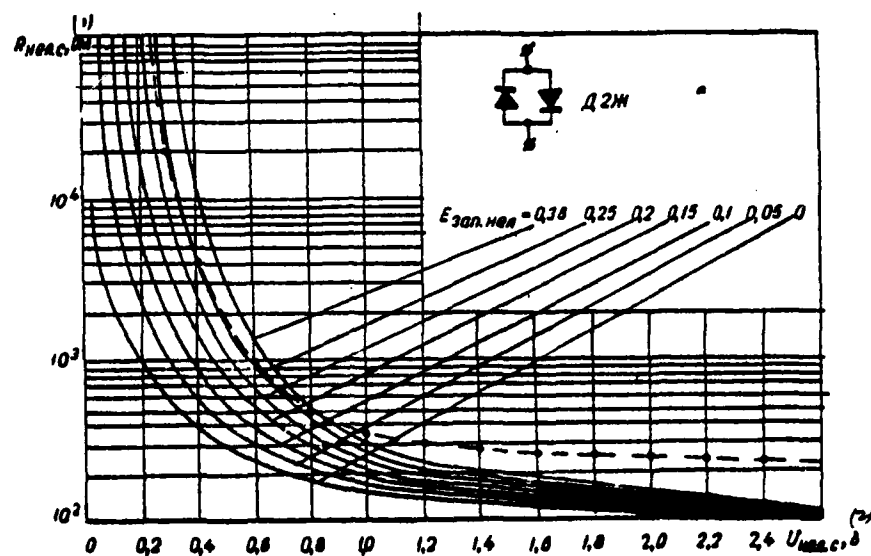


Fig. 100. Required and real curves of change of resisting nonlinear element/cell: --- $R_{\text{нел.с.}} = f(U_{\text{нел.с.}})$; — $R_{\text{нел.с.}} = f(U_{\text{нел.с.}})$.

Key: (1). ohm. (2). V.

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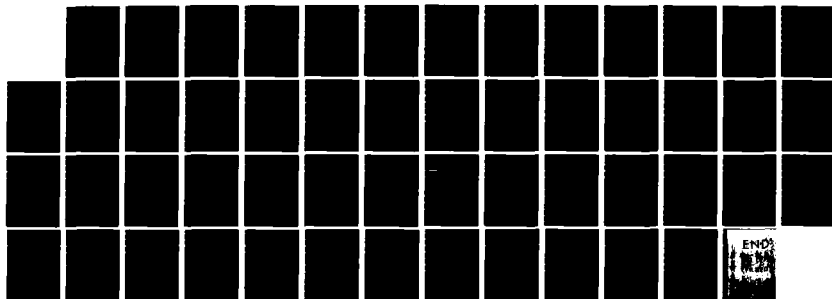
FUNCTIONAL ELECTRONIC AMPLIFIERS WITH BROAD DYNAMIC
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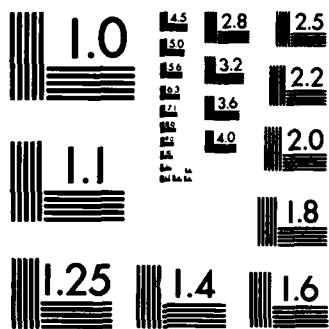
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7. From formulas, given in Table 3, we calculate that required AKh of nonlinear cascade/stage (dashed curve in Fig. 101).

According to general formula (2-11) with the use by curve $R_{\text{non. p}} + 100 \text{ ohm} = \varphi(U_m)$ we calculate the real AKh of latter/last (fourth) nonlinear cascade/stage. The calculation points of real amplitude characteristic are replaced in Fig. 101 and they coincide sufficiently well with the required characteristic. For the remaining nonlinear cascades/stages real AKh will have the same or even smaller divergences from the required characteristic, since curves

$R_{\text{non. p}} = \varphi(U_m)$ for these cascades/stages will more coincide precisely with required curve $R_{\text{non. o}} = \varphi(U_{\text{non. o}})$. Therefore for the calculation by the general/common/total AKh of amplifier it is possible to use an amplitude characteristic, only latter/last nonlinear cascade/stage.

8. From formula (1-34) we calculate precise LAKh (solid line in Fig. 102) of four-stage amplifier.

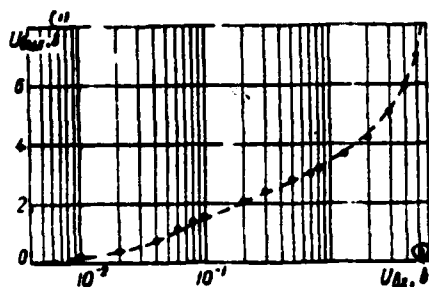


Fig. 101.

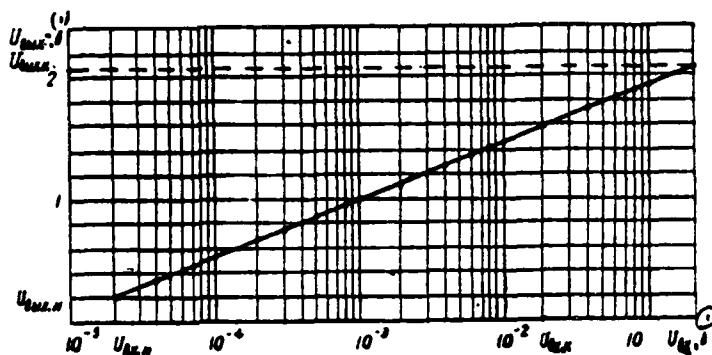


Fig. 102.

Fig. 101. Calculated and real amplitude characteristics of fourth nonlinear cascade/stage of logarithmic amplifier.

Key: (1). V .

Fig. 102. Calculated and real amplitude characteristics of four-stage logarithmic tuned amplifier.

Key: (1). V .

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From the real amplitude characteristics of nonlinear cascades/stages (in particular, on AKh the fourth nonlinear cascade/stage, depicted for Fig. 101) we design and construct

logarithmic amplitude characteristic UPCh, which consists of four nonlinear cascades/stages. The calculation points of true LAKh of UPCh are shown in Fig. 102. From the curves in Fig. 102, we see that real AKh in entire logarithmic range 80 dB differs from accurately logarithmic not more than to 3-4%.

With an example of the calculation of logarithmic video amplifier with the nonlinear feedback according to the cathode circuit it is possible to be introduced in work [7].

§ 2. Calculation of functional (logarithmic) amplifier with ARU according to the pentode grids of tubes.

For the amplifiers with ARU on the pentode grid, naturally, can be used the tubes, which have conclusion on the pentode grid. They include the pentodes of the type 6Zh1B, 6Zh2P, 6Zh4, 6Zh5B, 6Zh5P, 6Zh9P, 6P9 and so forth.

The distortions in the amplitude in the amplifier with the larger signals the less, the greater the extent of the grid-plate characteristic in the region of negative voltages/stresses on control electrode. From this point of view to the more advantageous side differ the tubes of the type 6P9, then - 6Zh4, 6Zh5B, 6Zh6P. In the tubes with the large slope/transconductance of the type 6Zh9P, 6Zh10P

and so forth the characteristic of tube is shorter, what is deficiency/lack.

During the design of amplifiers it is necessary to remember that the effect of adjustment the higher, the greater the cascades/stages included by adjustment.

Example of calculation. To calculate the schematic of logarithmic amplifier with ARU from the pentode grids of tubes, if are assigned the following parameters: $K_0=10^3$; $\Delta=4$ MHz; $f_0=30$ MHz; $D_{\max}=80$ dB. Logarithmic operation of signal according to the law of natural logarithm, i.e., $N=2.7(a=1)$.

The 1st stage. As a result of calculation are known following data: the type of tube 6Zh9P; the factor of amplification of cascade/stage with the work in linear conditions $K_n=10$; a number of cascades/stages $n=5$; voltage/stress on tubes $U_{av}=-1$ v.

The 2nd stage. 1. We choose diagram with one regulator (Fig. 45d).

2. According to characteristics of tube 6Zh9P when $U_n=-80$ v (Fig. 47b) we determine maximum permissible signal amplitude at input of latter/last cascade/stage. Taking into account that the maximum

controlling voltage/stress, which enters pentode grid U_p , there will be on the order of 80-100 V, maximum amplitude U_m can be taken order 2 V.

We accept $U_m = U_{m, max} (5) = 2$ V.

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3. Using formula (3-77), we calculate curve of dependence $\eta = \frac{U_p}{U_0} = f(X)$ with change X from 1 to 10^4 . Calculated curve 1 is shown in Fig. 103. From the figure one can see that η it varies from 0 to 0.75. U_0 - voltage/stress of the triggering/opening of tube on the pentode grid when $U_{cm} = -1$ V.

4. We determine minimum value of factor of amplification of cascades/stages at the end of dynamic range of LAKh, i.e., at the end of control

$$K_{min} = K_n = \sqrt{\frac{Z_n}{X_n}} = 10 \sqrt{\frac{10.2}{10^4}} = 2.5.$$

5. We determine maximum output voltage/stress $U_{max. n'}$ which corresponds to end/lead of LAKh,

$$U_{max. n} = U_{max. max} (5) = U_{ex. max} (5) K_{min} = 2 \cdot 2.5 = 5 \text{ V.}$$

6. We determine voltage/stress $U_{max. n'}$ with which begins with LAKh,

$$U_{\text{max. n}} = \frac{U_{\text{max. n}}}{Z_n} = \frac{5}{10,2} \approx 0,5 \text{ V.}$$

Along the axis of abscissas (Fig. 103) we plot/deposit the output voltage/stress of amplifier, which is the input voltage of regulator.

7. We choose diagram, which must manufacture controlling voltage/stress, which is changed according to the law $\eta=f(Z)$ (Fig. 103). Zaks' that required $\eta=f(Z)$ can be realized, if to the input of a tube of the type 6N9S when $U_a = 300 \text{ V}$ the bias voltage $U_{c\text{m}} = -1 \text{ v.}$ (Fig. 104) to supply from the output of amplifier negative voltage/stress with the transmission factor of detector $k_n = 0,70$, on which is fed the retarding voltage/stress $U_{a, n} = 0,5 \text{ V.}$ Value k_n is selected during the comparison of curve $\eta=f(Z)$ with the grid-plate characteristic.

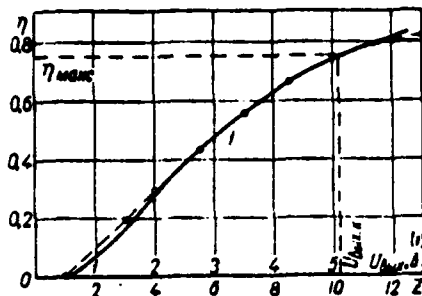


Fig. 103.

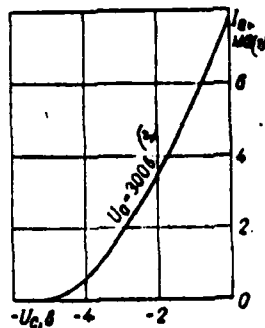


Fig. 104.

Fig. 103. Curves of dependence $\frac{U_p}{U_g} = f(Z)$ for five-stage amplifier with ARU on pentode grid.

Key: (1). V.

Fig. 104. Grid-plate characteristic 6N9S when $U_g = 300$ V.

Key: (1). mA. (2). V.

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Fig. 103 shows realizable law $\eta_p = \eta(Z)$ with the help of the tube 6N9S by points and it coincides sufficiently well with that required.

Certain difference in the required and realizable curves $\eta = f(Z)$

is observed in the beginning of the range of adjustment. However, this difference is easily removed by the selection of the corresponding mode/conditions of the work of the second amplifier of the regulator, which must manufacture the desired value of the controlling voltage/stress.

9. We determine voltage/stress of triggering/opening of tube 6Zh9P on pentode grid U_g . For this from the characteristics, depicted for Fig. 47, we design and construct the dependence of the slope/transconductance of tube on the voltage/stress on pentode grid $S=f(U_g)$ with different bias voltages U_{cm} (Fig. 105). For selected mode/conditions $U_{cm} = -1$ V straight line $S=f(U_g)$ we continue before the intersection with the axis of abscissas and we find $U_g = -107$ V. We accept $U_g = (-110$ V).

10. We determine maximum controlling voltage/stress $U_{p. max}$, which it must manufacture regulator

$$U_{p. max} = \eta_{max} U = 0.75 \cdot 110 = 82 \text{ V.}$$

Thus regulator must consist of detector with the transmission factor $k_d = 0.70$, on the preamplifier, which ensures the required law of a change in the controlling voltage/stress, and the fundamental amplifier, which ensures the required stress level U_p .

11. For purpose of uniformity for fundamental (second) amplifier of regulator we choose tube of type 6Zh9P. We accept the following mode of operation of tube (Fig. 47b); $U_{cm} = -2$ V; $I_{a.н} = 2$ mA (tubes L_1 in Fig. 60).

12. We are assigned by amplitude of change in anode current $\Delta I_a' = 22$ mA, which corresponds to change in voltage on input $u_{in}' = 1.5$ V.

13. We determine anodic lamp resistance of fundamental amplifier R_{a1} in Fig. 60).

$$R_{a1} = \frac{U_{p. max}}{\Delta I_a'} = \frac{82}{22 \cdot 10^{-3}} = 4 \cdot 10^3 \text{ ohm.}$$

According to GOST we choose $R_{a1} = 3.9$ kilohm.

14. We determine anodic lamp resistance of preliminary (first) amplifier. Since the tube 6N9S is twin triode, we connect in parallel both triodes. Then a maximum change in the anode current of the tube of the first amplifier according to Fig. 104

$$\Delta I_a' = 10.6 \text{ mA.}$$

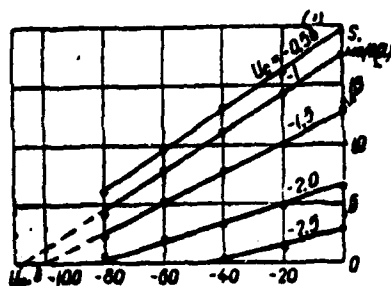


Fig. 105. Curves of the dependence of the slope/transconductance of tube 6Zh9P on the voltage/stress on the pentode grid.

Key: (1). V. (2). mA/V.

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Required anodic resistor/resistance (R_{a1} in Fig. 60)

$$R_{a1} = \frac{U_{\text{ex. p}}}{\Delta I_a} = \frac{1.5}{10.6 \cdot 10^{-3}} = 140 \text{ ohm.}$$

According to GOST we choose $R_{a1} = R_{a2} = 150 \text{ ohm.}$

The required nonlinearity of dependence $\eta=f(Z)$ in the beginning of adjustment (Fig. 103), as it is not difficult to see from the curves on (Fig. 47b), is obtained due to the nonlinearity of the grid-plate characteristic 6Zh9P of the second amplifier.

For obtaining the stable work of the second amplifier of regulator the voltage/stress on the screen grid of the tube of this amplifier is given from resistor/resistance of $R_{s,}$, over which the constant component of the cathode current of all tubes of amplifier stages flows/occurs/lasts. Since the cathode currents of tubes are constant (Fig. 47a), the voltage/stress, which is isolated on resistor/resistance of $R_{s,}$, is also constant.

The potentials of the cathodes of the tubes of amplifier stages and anode of the tube of the second regulator (L , in Fig. 60) are equal to each other, owing to which is provided zero potential on the pentode grids of the tubes of amplifier stages.

After the admission of the controlling voltage on the input of regulator the anode current of tube L , increases according to the law $\eta=f(Z)$ and the potential of the pentode grids of tubes UK is reduced according to the same law. As a result the necessary gain control in cascades/stages and $LAKh$ of multistage amplifier is realized.

The experimental amplifier, made on the diagram in Fig. 60, had data, indicated into § 2, Chapter 3. The real characteristic was somewhat different from the calculated to the smaller side at the end of the adjustment. This is explained by certain overloading of latter/last cascade/stage.

§ 3. Calculation of logarithmic transistorized amplifier with the addition of the detected voltages/stresses.

The calculation of logarithmic amplifier in this case is produced into two stages. During the selection of amplifier circuit it is necessary to be guided by the following:

transistors in amplifier stages must work without the overloading;

diagram must be simple in the tuning/adjusting in the implementation of LAKh of high accuracy;

the recurrence of the parameters of amplifiers from one copy to the next must be good.

To the greatest extent the diagram in Fig. 71, described in Chapter 4, satisfies these requirements. Let us consider the calculation of this diagram.

Example of calculation. To calculate logarithmic UPCh with the video output, whose diagram is depicted in Fig. 71, if are assigned the following parameters: $K_0=3 \cdot 10$; $f_0=30$ MHz; $U=4$ MHz; $D_{\text{MAX}}=75$ dB; $\sigma=120$ mV/dB.

1-stage. As a result of the calculation, known are the following data necessary for calculation of the amplifier in the nonlinear mode: type of transistor - P417; operating mode of the transistor $I_{\text{a. n. n}}=4$ mA; amplification factor of the cascade in the linear mode $K_n=4.5$; coefficient of inclusion of the circuit $m=0.2$; number of cascades $n=7$; value of resistance of ARU in circuit of emitter $R_e=R_{e1}+R_{e2}=2.7$

k Ω . With the selected mode and resistance R_e fulfilled is the inequality

$$I_{\text{a. n. n}} R_e > U_{\text{a. d. n.}}$$

Because of this, for the calculation of AKh of the cascade, it is possible to use formula (3-70).

2-stage. Calculation of amplifier in the nonlinear mode/conditions.

1. According to experimental input characteristic (Fig. 106) we determine coefficient γ . For this we are assigned $I_{a1}=8$ μ A and $I_{a2}=10$ μ A. From the curve on Fig. 106 we find $U_{a1}=0.38$ V; $U_{a2}=0.28$ V.

From the relationship/ratio

$$\frac{i_{01}}{i_{02}} = e^{\gamma(U_{a, 01} - U_{a, 02})} = \frac{80}{10}$$

we determine $\gamma \approx 15$.

2. From formula (3-70) we calculate AKh of cascade/stage (curve 1 in Fig. 107) and also voltage in circuit $U_{\text{max}} = I(U_{\text{ax max}})$ (curve 2 in Fig. 107)

$$U_{\text{max}} = \frac{U_{\text{ax max}}}{n}.$$

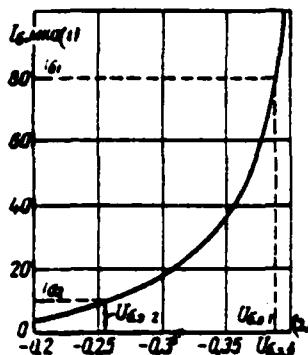


Fig. 106.

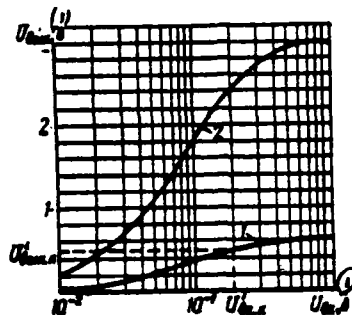


Fig. 107.

Fig. 106. Experimental input characteristic of transistor of type P-417.

Key: (1). μA . (2). V.

Fig. 107. Amplitude characteristics of amplifier stage: 1 - on output; 2 - on duct/contour.

Key: (1). V.

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3. We design detector for transistor of type P416, which has parameters: $f_a = 120$ MHz; $r_o = 80$ ohm; slope/transconductance at zero frequency $S_o = 140$ mA/V.

4. We determine modulus/module of mutual conductance of transistor P416 at frequency of rectified signal $f_r = 30$ MHz:

$$|S| = \frac{S_0}{\sqrt{1 + (\omega\tau)^2}} = \frac{140}{1 + \left(2\pi \cdot 30 \cdot 10^6 \cdot \frac{140 \cdot 10^{-8} \cdot 80}{2\pi \cdot 120 \cdot 10^6}\right)^2} \approx 47 \text{ mA/V}$$

where $\tau = \frac{S_0}{2\pi f_a}$.

5. We determine mutual conductance of detectors taking into account action of feedback due to control resistors $R_p (R_{20} - R_{20})$, connected in emitter circuit of transistors. Resistors/resistances R_p take 20-40 ohms. We accept $R_p = 30$ ohm. Then according to work [22]

$$S_x = \frac{S}{1 + R_p(S + g)} \approx \frac{S}{1 + S_p} = \frac{47 \cdot 10^{-3}}{1 + 47 \cdot 10^{-3} \cdot 30} = 19 \text{ mA/V.}$$

since $S \gg g$. We accept $S_x = 20$ mA/V.

6. We choose mode/conditions of work of transistors of detection cascades/stages with low currents of collectors/receptacles. Then angle of cutoff θ is close to $\pi/2$. Since the resistors/resistances in the emitter circuits are blocked by the capacitors/condensers of sufficiently great capacity, voltage/stress on them for the pulse action time substantially is not changed. Consequently, it is possible to consider angle of cutoff θ as constant.

7. Choosing different values of coefficient of connection of detector to duct/contour m_x , it is possible to change value of voltage of signal, which enters detector. With an increase in m_x the voltage on the input of detector and the slope/transconductance of LAKh of amplifier increase. However, with increase in m_x the effect of the parameters of detector on the oscillatory circuit increases. We accept $m_x = m = 0.2$. Then the output voltage/stress of cascade/stage (curve 1 in Fig. 107) is the input voltage of detector, i.e.

$$u_{\text{max. 1}} = u_{\text{ss. DET}}$$

8. For exception/elimination of effect of parameters of transistors of amplifier stages on maximum output voltage/stress $u_{\text{max. 1}}$ and, therefore, and to accuracy of LAKh of amplifier, detectors we place in mode/conditions of limitation on maximum. The threshold of limitation is selected $u_{\text{max. 1}} = 0.5 \text{ V}$ (Fig. 107), which corresponds to voltage on the input of cascade/stage $u_{\text{ss. 1}} = 0.2 \text{ V}$.

9. We determine maximum amplitude of rectified current of collector/receptacle with $u_{\text{max. 1}}$

$$I_{\text{L. MAX}} = c_p S_A u_{\text{max. 1}} = 0.319 \cdot 20 \cdot 10^{-4} \cdot 0.5 = 3.2 \text{ mA}$$

where

$\alpha=0.319$ with $\theta=\pi/2$.

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10. We determine total resistance of load of detector made of following considerations. Upon transfer from initial operating point $I_n \approx 0$ into the mode/conditions of limitation the voltage/stress on the collector/receptacle varies from supply voltage E_n to zero. We accept $E_n = 12$ V. Then

$$R_{n, \lambda} = \frac{E_n}{I_n} = \frac{12}{3.2} = 3.75 \cdot 10^3 \text{ ohm.}$$

Resistor/resistance $R_{n, \lambda}$ is equal to the sum of the resistors/resistances

$$R_{n, \lambda} = R_{n, \lambda} + R_3 = R_{n, \lambda} + R_{n, 0}.$$

11. We assume that all cascades/stages identical. Then the working dynamic range of cascade/stage d (according to § 5 of Chapter 1) is equal to his amplification factor with the work in the linear conditions, i.e.

$$d = K_n = 4.5 \text{ OR } d_{(00)} = K_{n(00)} = 13 \text{ dB.}$$

12. Using expression (1-142), we design necessary incremental stress for summator (resistor/resistance $R_{n, 0}$) due to one nonlinear

cascade/stage

$$\Delta U = \Delta \varphi = \sigma \ln d = 120 \cdot 10^{-3} \cdot 13 = 1.55 \text{ V}.$$

13. We calculate resistor/resistance of summator $R_{\Sigma 0} = R_1$ on the basis of coefficient of division in load circuit of detector

$$k_{\text{det}} = \frac{R_1}{R_{\Sigma A}} = \frac{\Delta U}{E_N}.$$

Whence

$$R_1 = R_{\Sigma A} \frac{\Delta U}{E_N} = \frac{3.75 \cdot 1.55}{12} = 480 \text{ ohm}.$$

Then resistor/resistance $R_{\Sigma A}$ in the circuit of the collector/receptacle

$$R_{\Sigma A} = R_{\Sigma 0} - R_1 = 3.75 - 0.48 \approx 2.95 \text{ kilohm}.$$

In this case relation $\frac{R_{\Sigma A}}{R_1} = \frac{2.95}{0.48} \approx 6 < 10$ (condition 4-17) is performed insufficiently, i.e., sufficiently precise LAKh with the slope/transconductance $\sigma = 120 \text{ mV/dB}$ with relieving of output voltage directly from the summator cannot be obtained.

14. We choose according to GOST $R_2 = R_{\Sigma 0} = 100 \text{ ohm}$; $R_{\Sigma A} = 3.6 \text{ kilohm}$.

Then

$$\sigma_1 = \frac{E_N R_2}{R_{\Sigma A} \ln d} = \frac{12 \cdot 0.1}{3.7 \cdot 13} = 2.5 \text{ mV/dB}.$$

15. For obtaining required slope/transconductance of LAKh to output it is necessary to connect linear amplifier with factor of amplification

$$K_{\text{LAKh}} = \frac{e}{e_1} = \frac{120}{25} = 4.8.$$

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In the diagram Fig. 71 video amplifier is assembled on transistor T, of the type P416. For the linearization of AKh of video amplifier with the emitter circuit of transistor is connected the sufficiently high resistor/resistance of feedback R_{L} , shunted by a small capacity/capacitance, which corrects the frequency characteristic of the video amplifier in the region of the highest frequencies.

16 ¹.

FOOTNOTE ¹. If detectors work without the mode/conditions of limitation, it is expedient to solve inverse problem, namely: first from formulas (1-143) and (1-144) to calculate values k_{LAKh} , and then taking into account condition (4-17) to determine values R_{LAKh} , R_{L} and S_{LAKh} . ENDFOOTNOTE.

According to simplified formulas (1-143), (1-144) taking into account

new values σ_1 and $u_{\text{BEMX}, N}$ we calculate the transmission factors of 1st and all remaining detectors

$$k_{n1} = \frac{\sigma_1 \ln d}{u'_{\text{BEMX}, N} \left(1 - \frac{1}{K_N}\right)} = \frac{25 \cdot 10^{-3} \cdot 13}{0,5 \left(1 - \frac{1}{4,5}\right)} = 0,83;$$

$$k_{n2} = k_{n3} = \dots = k_{ni} = k_{ni} = \frac{\sigma \ln d}{u'_{\text{BEMX}, N}} = \frac{2,5 \cdot 10^{-3} \cdot 13}{0,5} = 0,65.$$

As it is not difficult to see, values $k_{n1} = 0,65$ in the diagram are already realized since slope/transconductance σ_1 it is designed, on the basis of the values S_{n1}, R_{n1}, R_1 and $u'_{\text{BEMX}, N}$ accepted.

It is most easy to obtain value $k_{n1} = 0,83$ increasing mutual conductance of 1st detector S_{n1} and reducing resistor/resistance $R_{p1} (R_{n1})$ in the emitter circuit of transistor T_1 . We find desired values S_{n1} and R_{p1} :

$$S_{n1} = S_{n1} \frac{k_{n1}}{k_{n1}} = 20 \cdot \frac{0,83}{0,65} = 25 \text{ mA/V};$$

$$R_{p1} = \frac{S - S_{n1}}{S S_{n1}} = \frac{47 - 25}{47 \cdot 25} = 19 \text{ ohm.}^{(2)}$$

Key: (1). mA/V. (2). ohm.

We accept $R_{p1} (R_{n1}) = 20 \text{ ohm}$.

17. According to amplitude characteristics of cascades/stages, whose ordinates are multiplied by coefficients k_{n1} and $K_{n, \gamma}$, i.e.

$$U_{\text{BEMX}, i} = u_{\text{BEMX}} k_{n1} K_{n, \gamma}$$

taking into account mode/conditions of limitation in detectors we

construct AKh of amplifier (Fig. 108). For this we represent AKh of cascades/stages, beginning from the latter (curve 7 in Fig. 108), in the system of coordinates $U_{\text{out}} = f(a)$ of that calibrated of the input and output voltages of amplifier. The amplitude characteristics of cascades/stages 6, 5, 4, 3 and 2 are constructed by the simple transfer of characteristic 7 along the axis of abscissas to the value of the working dynamic range of cascade/stage, i.e., on $d=13$ dB. In Fig. 107 $a = 20 \lg \frac{U_{\text{sa}}}{U_{\text{sa},0}}$, where $U_{\text{sa},0} = 2$ v - zero reference level. Summarizing the ordinates of curves 1-7, we obtain AKh of amplifier (curve 8).

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18. We carry out straight line 9, which corresponds to precise LAKh, calculated by formula (1-34), and we determine dynamic range, in which objective parameter sufficiently coincides precisely with precise. In Fig. 107 it is evident that $a_n = 22$ dB; $a_n = 105$ dB; i.e.

$$D_{\text{MAX}} = 83 \text{ dB.}$$

Maximally expected errors of LAKh, which will be obtained on the assumption that the cascades/stages work in the mode/conditions of linear amplification - limitation, they can be calculated by formulas (1-184), (1-186). Calculated errors will be more than real ones, since cascades/stages work in the nonlinear modes/conditions.

19. Dynamic range of LAKh can be increased on 10-13 dB, if we to input of amplifier connect further (zero) detector with transmission factor $k_{zo} = 0.83$. In this case the first detector must have a transmission factor $k_{z1} = 0.65$. In Fig. 71 the null detector is assembled on transistor T_1 .

Amplitude characteristics on the output of the 1st and null detectors for the present instance Fig. 107 depicts as prime and are designated through 1* and 0. Fig. 107 depicts the amplitude characteristic of amplifier with the zero (further) detector as curved 10. In the same figure (dashed curve 11) it is shown the AKh of the amplifier without the linear video amplifier with relieving of output stress directly from the summator (resistor/resistance R_{11}).

The mock-up, made on the diagram (Fig. 71), tested in the laboratory. Experimental data (point in Fig. 108) very well coincide with the calculated ones.

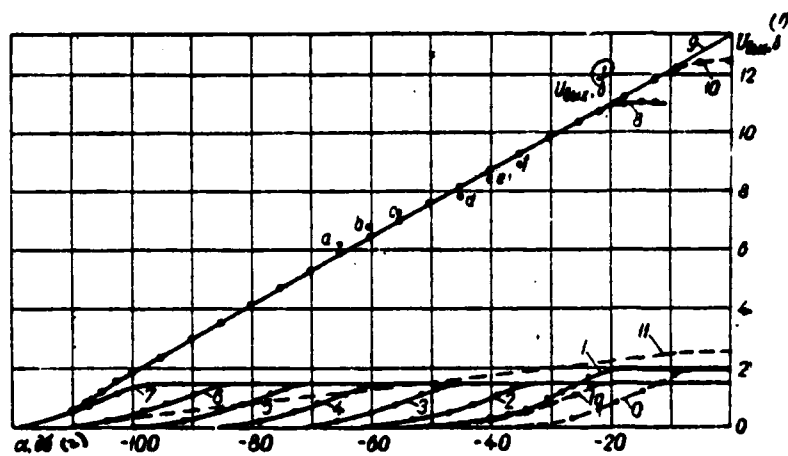


Fig. 108. Amplitude characteristic of logarithmic transistorized amplifier with the addition of voltages/stresses.

Key: (1). V. (2). db.

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54. Tuning of functional amplifier.

The adjustment of amplifier with FAKh, just as calculation, should be carried out in two stages:

1st stage. Regulation and tuning/adjusting in linear mode/conditions.

2nd stage. Adjustment of amplifier for obtaining precise FAKh.

If FAKh in n-cascade amplifier is realized according to the method of changing the factor of amplification of nonlinear cascades/stages, the adjustment of amplifier is begun from the adjustment of separately each nonlinear cascade/stage. In this case it is necessary that the amplitude characteristics of nonlinear cascades/stages would correspond precisely to calculated ones. Special attention should be paid to satisfaction of working conditions for the successive of nonlinear cascades/stages.

Only in this case it is possible to sufficiently easily and rapidly obtain precise FAKh of amplifier as a whole. After all nonlinear cascades/stages are controlled, is regulated logarithmic amplifier as a whole.

If FAKh in the amplifier is realized according to the method of adding the output effects, special attention should be paid for the selection of the transmission factors of the corrective elements/cells (detectors, correcting cascades/stages). Methodology of the selection of the transmission factors of detectors let us examine based on the example Fig. 108. After plotting of amplitude curve/characteristic of amplifier, are determined the sections of dynamic range, in which LAKh is formed/shaped with some cascade/stage. Let us suppose experimental of characteristic differs from precise to the large side at points a, b, c, d, e and f (Fig. 108). It is obvious that for eliminating the error of LAKh it is necessary and the first turn to decrease the transmission factor of the 3rd detector and to increase the 2nd. After this to again plot the curve of amplifier. If there will again be divergences, then by change of the transmission factors of detector reduce them.

In order to eliminate the effect of the characteristics of the previous cascades/stages on the characteristic of amplifier during the adjustment of the i cascade/stage, it is necessary to regulate

(to change the transmission factors of detectors) amplifier, by beginning from the latter to by ending with the first (zero) detector.

The amplitude characteristic of separate selective nonlinear cascade/stage and entire amplifier as a whole it is possible to shine sufficiently accurately on the block diagram, depicted in Fig. 109.

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As the source of signal can be used high-stability standard signal generator (GSS), from output of which it is possible to remove/take usual sinusoidal high-frequency oscillations and oscillations, modulated in the amplitude by impulses/momenta/pulses or audio frequency. The high stability of amplitude and depth of modulation of the generatable oscillations is the fundamental requirement, presented to the signal generator.

The amplitude of the oscillations, removed with GSS, can be insufficient; therefore after signal generator it is necessary to switch on amplifier with strictly constant coefficient of amplification in time. The value of the signal, which enters the input of adjustable logarithmic amplifier (LU), it is possible to set with an attenuator which stands at the output of the signal generator. Since the attenuators of signal generators they frequently have

large error in the calibration, for removing/taking AKh the amplifier with the high accuracy it is necessary between the amplifier U₁ and tuned LU to include/connect the further broadband attenuator A, by using which, it is possible with the high accuracy to change attenuation within the limits from 0 to 100-120 dB.

For the accuracy of the removal/taking of LAKh of amplifier it is necessary that with the attenuator it would be possible to change attenuation spasmodically through 3-5 dB with the accuracy not less than 1%. Diagram and construction/design of this attenuator is described in work [41]. During the use of a further precise attenuator the attenuation in the attenuator of signal generator can be reduced to zero.

As the measuring meter I, connected at the output of logarithmic amplifier, it is possible to use a cathode voltmeter of the high class of precision. During rough plotting of amplitude curve/characteristic it is possible to use an oscillograph of the type 25I or S1-8 (UO-1M).

The amplitude curve of cascade/stage or amplifier is plotted as follows. At the output of GSS maximum voltage/stress is installed, and in the attenuator A they introduce in the complete attenuation. If output of voltage/stress GSS is small, then is connected amplifier

U₁ (key/wrench P₁ is connected). Then, gradually decreasing the attenuation, introduced by attenuator, with cathode voltmeter is fixed/recorded output voltage/stress for each value of the attenuation of attenuator.

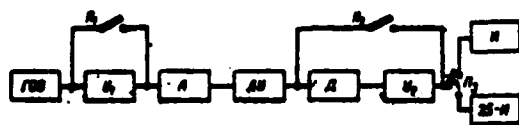


Fig. 109. Block diagram of installation/setting up for plotting the amplitude curve/characteristic of logarithmic amplifier.

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They reduce the attenuation of attenuator until to input of LU begin to enter the signals, which emerge from the dynamic range of LAKh of amplifier. After plotting points to the semilog diagram, is obtained real AKh of amplifier, which with the small divergences from the accurately logarithmic amplitude characteristic must be straight line.

An error in the real AKh of amplifier is determined by the divergence of points from straight line. An error in the removal/taking of AKh amplifier is determined by the errors, introduced by attenuator, measuring meter and subjective errors of operator.

For the rapid testing of the accuracy of real LAKh of selective amplifier it is possible to apply the method of the modulated

oscillations, whose essence consists of the following. If we to the input of logarithmic amplifier feed high-frequency oscillations with an amplitude of U_m , modulated in the amplitude by audio frequency with the modulation factor $m = \frac{U_a}{U_m}$, where U_a — amplitude of the enveloping audio frequency at the input of amplifier, then at the output of the amplifier of amplitude of the enveloping audio frequency

$$U_{a \text{ max}} = K_a U_{aL} n a \ln(1 + m) \quad (6-7)$$

and with $m = \text{const}$ it will not depend on the value of the amplitude of high-frequency oscillation at the input of amplifier in entire dynamic range of LAKh.

Expression (6-7) can be registered thus:

$$U_{a \text{ max}} = \sigma_N \ln(1 + m). \quad (6-8)$$

Whence the slope/transconductance of LAKh of the amplifier

$$\sigma_N = \frac{U_{a \text{ max}}}{\ln(1 + m)}. \quad (6-9)$$

For the more precision determination of the value of slope/transconductance σ and to this point of dynamic range LAKh coefficient m is taken not more than 0.02-0.05.

In this case formula (6-9) takes the form

$$\sigma_N = \frac{U_{a \text{ max}}}{m}. \quad (6-10)$$

If we output potential of selective amplifier linearly rectify

with the help of the detector D and to isolate the envelope of audio frequency, then the amplitude of envelope can be accurately measured by cathode voltmeter (amplifier U_1 with the coefficient K , it can be connected for an increase in the amplitude of detected envelope at the low values of m, K , and U_{max}).

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Taking into account coefficient K ,

$$\sigma_N = \frac{U_{\text{max}}}{K \cdot m}. \quad (6-11)$$

The accuracy of LAKh of selective amplifier is rapidly checked as follows. In standard signal generator the specific modulation factor m is installed. The scale of voltmeter is graduated in accordance with equality (6-10) or (6-11). The attenuation of attenuator further is changed and is observed the arrow/pointer of voltmeter. In the case of precise LAKh the arrow/pointer of the voltmeter, connected after amplifier U_1 , is motionless and shows value (6-11) with a change in the voltage on the input of amplifier in entire dynamic range of LAKh. If pointer of voltmeter at some values of input voltage differs from the value, determined by expression (6-11), then this indicates that with the given input voltage the objective parameter of amplifier differs from accurately logarithmic. If the arrow/pointer of voltmeter differs from the

assigned magnitude to the large side, this means that the slope/transconductance of real AKh increases (logarithm to the base N is reduced) in comparison with the given one and, vice versa.

By the method of the modulated oscillations it is possible very rapidly to check AKh of amplifier and to come to light/detect/expose the sections of characteristic, which deviate from the accurately logarithmic.

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appendices.

Appendix 1.

Analytical expressions of Y-parameters for the transistors:

$$Y = \frac{1}{r_0} \cdot \frac{S r_0 + j\omega\tau}{1 + j\omega\tau}; \quad S = \frac{S_0}{2 + j\omega\tau};$$

$$Y_{oop} = \frac{S_{oop} + j\omega C_{0. n}}{1 + j\omega\tau}; \quad Y_i = \frac{1}{R_i} + j\omega \frac{S_0 r_0 C_{0. n}}{1 + j\omega\tau} + j\omega C_{0. n},$$

where

$$S_0 = \frac{G_0'}{1 + (S_{0. o} + S_{0. n}) r_0}; \quad \tau = \frac{(C_{0. o} + C_{0. n}) r_0}{1 + (S_{0. o} + S_{0. n}) r_0}.$$

The cut-off frequencies:

for the common-base circuit

$$f_a = \frac{S_0 r_0}{2\pi\tau};$$

for the common-emitter connection

$$f_p = \frac{S_0 r_0}{2\pi\tau}.$$

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Appendix 2.

Upon transfer from r-parameters into Y-parameters of the T-shaped equivalent schematic of the transistor, connected on the common-base circuit,

$$S_0 = \frac{\alpha_0}{(1-\alpha_0)r_0 + r_e}; \quad \mu = \frac{\alpha_0 r_K}{r_0 + r_e}; \quad D = \frac{1}{\mu};$$

$$R_i = \frac{\mu}{S_0}; \quad g = S_0 \frac{1-\alpha_0}{\alpha_0}; \quad g_{oop} = \frac{S_0 r_0 r_e}{r_K(r_0 + r_e)};$$

$$r_0 = r_e + r_{e1}; \quad \tau = \frac{S_0 r_0}{2\pi f_e}.$$

where α - coefficient of current amplification of emitter; r_e - resistor/resistance of emitter; r_0 - resistor/resistance of base; r_K - resistor/resistance of collector/receptacle.

Upon transfer from h-parameters, measured in the diagram with common emitter,

$$g = \frac{1}{h_{11}'}; \quad g_{oop} = \frac{h_{12}'}{h_{11}'};$$

$$S_0 = \frac{h_{21}'}{h_{11}'}; \quad \frac{1}{R_i} = h_{11}' - \frac{h_{12}'}{h_{21}'} h_{22}'.$$

Appendix 3.

Coefficients A and B must be selected in such a way that the functions $F(x)$ and $f(x)$ would coincide the minimum at three points, namely: $x=1$, 0, 5 and 0. Coincidence of $f(x)$ and $F(x)$ even at three points provides a sufficient accuracy of calculations in view of the simple character of function $f(x)$.

It is easy to see that functions $F(x)$ and $f(x)$ become zero with $x=1$. Thus, $F(x)$ and $f(x)$ coincide with $x=1$.

In order to obtain equality $F(0)=f(0)$ with $x=0$, coefficient A , according to (5-121), it must be equal to

$$A = f(0) = f(x)_{x=0} \quad (3-1)$$

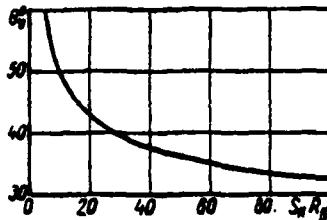


Fig. 110.

Fig. 110. Curve of dependence $\theta_y = S_n R_n$.

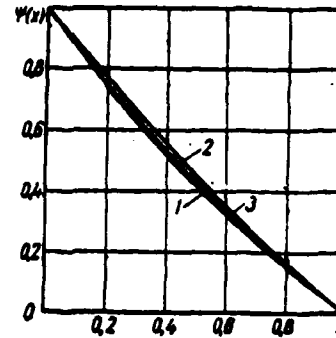


Fig. 111.

Fig. 111. Curves of dependences $\psi(x)$ at different values $S_n R_n$:

1 - $S_n R_n = 10$; 2 - $S_n R_n = 10$; 3 - $S_n R_n = 10$.

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We find value of $f(0)$.

Substituting the value of x from expression (5-117) in (5-119), we obtain

$$f(x) = \frac{S_n}{\pi C_n} (\lg \theta - \theta) \frac{\cos \theta}{\cos \theta_x} - \frac{\cos \theta}{\cos \theta_x} \cdot \frac{1}{C_n R_n}. \quad (3-2)$$

During first periodine voltage/stress U_n on resistor/resistance R_n is small (considerably less than the amplitude U_m) and therefore

angle of cutoff θ , is maximum and equal to approximately $\pi/2$. Equalizing x to zero in expression (3-2) and substituting the value $\theta = \theta_0 = \pi/2$, we obtain the following expression for A

$$A = f(0) = \frac{S_N (\sin \theta - \theta \cos \theta)}{\pi \cos \theta_x} - \frac{\cos \theta}{\cos \theta_x} \cdot \frac{1}{C_N K_N} = \frac{S_N}{\pi C_N \cos \theta_x}. \quad (3-3)$$

Formula for calculating the coefficient of B is obtained, on the basis of the requirement of coincidence $F(x)$ and $f(x)$ with $x=0.5$.

$$f(0.5) = M f(0). \quad (3-4)$$

The value of coefficient of M is easy to find, after constructing plotted function $f(x)$. By given value $S_N R_N$ the angle of cutoff in the steady-state mode/conditions (Fig. 110) is determined. They are further assigned by the position of the values of angle θ in limits of 90° to the values, close to θ_0 , and in formulas (5-117), (5-119) are determined the appropriate values of x and $f(x)$.

Equalizing $F(0.5)$ and $f(0.5)$ to each other and taking into account relationships (5-121), (3-1) and (3-4), we obtain

$$M = 0.5(1 - 0.5B). \quad (3-5)$$

Whence

$$B = 2(1 + 2M). \quad (3-6)$$

Curves $\psi(x) = f(x)/f(0)$, necessary for determining the coefficient of M , they are depicted in Fig. 111.

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